

WHEN LIGHTNING STRIKES...

Lightning strikes can be a problem for your computer.

Herb Friedman

■ Before the age of electronics, the fear of lightning was not so much its immense electrical energy but the fire caused by the "hit" or "strike." Lightning protection meant some means of directing lightning to where it wanted to go—to ground. This was done by installing grounded metal rods—lightning rods—on top of houses, barns, windmills, whatever. The purpose was twofold. It grounded the energy and it conducted the energy around the structure.

With the electronic age, we inherited the worry of the electronic field produced by the strike. Very high fields are built up in the ground during an electrical storm. The ground field can track under a charged cloud and travel for miles before anything occurs. When the charge between the ground and the cloud

becomes sufficient the energies attract and we get the lightning strike. At the moment of strike there is an enormous discharge of ground energy, which creates a rapid expansion and collapse of a local electric field. Any collapsing or expanding field induces both current and voltage in wires. Although lightning might strike a tree in front of a house, the energy field produced during the strike can induce high frequency transient voltages in antenna lead-ins, power lines and telephone lines located 50 to 100 feet or more from the actual hit.

Lightning arrestors

While a lightning arrestor can protect antenna lead-in wires from direct hits and devices such as Zener diodes and MOV's (Metal Oxide Varistors) can protect conventional electronic equipment against powerline surges, microprocessors and other electronic devices are often "blown" by transients induced by the lightning strike which are not of sufficient power to trigger a lightning arrestor, but which are strong enough to blow discrete components. There are many instances on record where the voltage surge induced in electric, telephone and antenna lead-in wires by a local lightning hit has been known to zip through the wiring and blow a TV tuner or the microprocessor of a personal computer, even if the computer is turned off or disconnected from the power line.

Even if disconnected? Yes! Solid state devices—particularly microprocessors—are sensitive to external high-frequency voltage surges because the surge can enter the equipment through any external wire connection, not just the power line. While low and medium frequency AC can be stopped by an open switch, an RF choke or a small reactance, high-frequency energy, such as that produced by a lightning strike, can leap across open switches, couple from one wire to another through otherwise insignificant capacitive coupling, even pass between insulated transformer windings. While we have made great strides in squashing power line transient surges with devices such as the MOV and gas-discharge tube, technology has not caught up with the problems caused by lightning surges in personal computer equipment.

For example, electrical storms are unusual in my locality, we get one or two every five years or so. Between the previous one and this year's electrical storm, several computers and modems have taken up

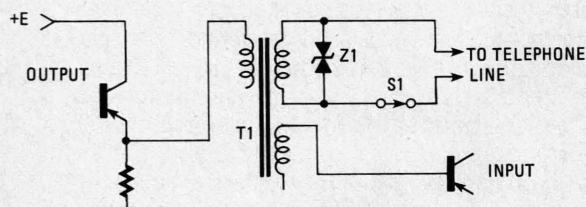


FIG. 1—THE BASIC ELEMENTS of a direct-connect modem's interface to the telephone line. Z1 is a surge suppressor that squashes transient voltages appearing across the line to approximately 50 volts. S1, which disconnects the modem from the line, is usually SPST.

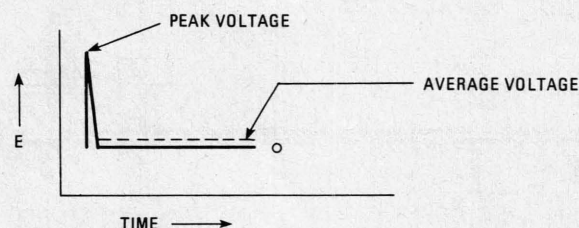


FIG. 2—THE LOCAL GROUND-DISCHARGE FIELD energy is almost an instantaneous rise and decay. This produces high-frequency energy through the upper VHF frequencies. In practical terms, it is high-voltage RFI.

residence in my home. In our most-recent electrical storm, lightning struck outside and disintegrated ten feet of curb. That was the only damage I thought, because my two computers were switched off, as were their modems. However, when I attempted to access an on-line information service, I discovered that both modems were blown, as was one computer's RS-232 driver (the one used for the modem). Coincidentally, the same storm passed through the area 20 miles away where my office is located, and lightning had struck the grounded radio/TV tower on the building.

Since my regular modems were inoperative, I borrowed the office modem which had been switched off during the storm, only to find that it too, was "blown."

Interesting situation

Three modems of different design and manufacture had been blown during an electrical storm. All had some form of voltage surge protection, all had been switched off and one didn't even have a connection to the power line because it is powered directly from the phone line. A fourth modem, built into an IBM-PC which was connected on-line because it serves as a "host" for a voice message center, wasn't damaged at all!

What to do

As near as we can determine, the three "blown" modems were damaged because of the way they were switched to the telephone line. Figure 1 is a simplified diagram of a direct-connect modem interface. Transformer T1 is the coupling transformer (repeat coil), Z1 is the telephone line overvoltage or surge-protection device, and S1 is a SPST switch that connects the modem to the telephone line. Normally, a voltage surge on the line is squashed by transient suppressor Z1, which might be a MOV, Zener diode, or gas-discharge tube. Usually, Z1 limits the surge to about 50 volts.

As shown in Figure 2, a voltage developed in the telephone line by a lightning strike in the general area

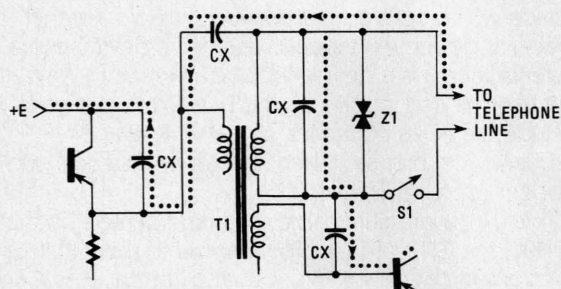


FIG. 3—CAPACITOR "Cx" IS THE NORMAL circuit capacity between transformer windings, solid-state device elements and component interconnecting wires. Cx is usually so small that it doesn't exist at audio frequencies but can represent a low impedance or "dead short" at the high frequencies of a lightning strike. When S1 is opened the lightning strike energy can follow the dotted path into the modem and eventually the high voltage bus where it is distributed throughout the modem, even to the modem's special driver.

will be a steep pulse. Although the average voltage—meaning the peak voltage averaged over a period of time—might be relatively low and safe, the peak value might be several hundred volts, and it is predominantly high frequency—extending well into the upper VHF range. High frequencies don't need a direct connection to pass from one wire to another. What is otherwise a minute capacity between wires and circuits, can appear to be "dead short" to high frequencies.

Imagine that the ground discharge of a lightning strike induces the transient waveform shown in Figure 2 on the telephone line connected to the modem circuit shown in Figure 3. Although S1 is open, disconnecting the modem from the telephone line, one wire is still connected to the line itself because almost all modems switch only one side of the telephone connection. We now have a high-frequency voltage surge on the telephone line which is looking for a path to ground. As shown by the dotted line in Figure 3, it finds the path to ground through the components in what we believe to be a disconnected modem. Notice that the energy represented by the dotted line flows through the capacity between T1's windings. The capacity might be very small, but it is there. To normal speech frequencies the capacity represents essentially an infinite impedance, but to the high frequency component of a lightning-caused transient surge the small capacity represents a low impedance, or even a "dead short," so the spike produced by the voltage surge passes through T1's secondary (telephone side) to the primary (modem side) to the modem's voltage supply bus. Now we have a high-voltage transient on what is usually a five- to nine-volt power bus. The result? ZAP!!! A fistful of blown IC's. If the surge is sufficiently high, it can ride through a blown IC in the modem and into the RS-232 driver, ZAPPING another handful of parts.

You might ask why Z1, the modem's internal transient suppressor(s) didn't stop the surge. Refer back to Figures 1 and 3. Z1 is connected across the telephone line: It is supposed to clamp surges appearing across the line, between both wires. But when switch S1 is opened to disconnect the modem from the telephone line, the transient suppressor has no effect on anything in the circuit. Z1 provides protection only when S1 is closed and the modem is connected to the line. This is probably why the direct-connect modem, connected on-line during the storm, wasn't damaged. Essentially, the risk of damage during an electric storm is increased when the modem is switched off the line.

My "cure"

Most likely, modems should be manufactured with a double pole output switch so that both sides of the telephone line can be disconnected from the modem. Until such time as the manufacturers get around to changing their designs, I have installed a DPST knife switch on each telephone line to make certain the modems are really disconnected from the line. The inconvenience of "throwing the switch" is a lot less costly than the charges to repair or replace three modems and an RS-232 port. ◀▶

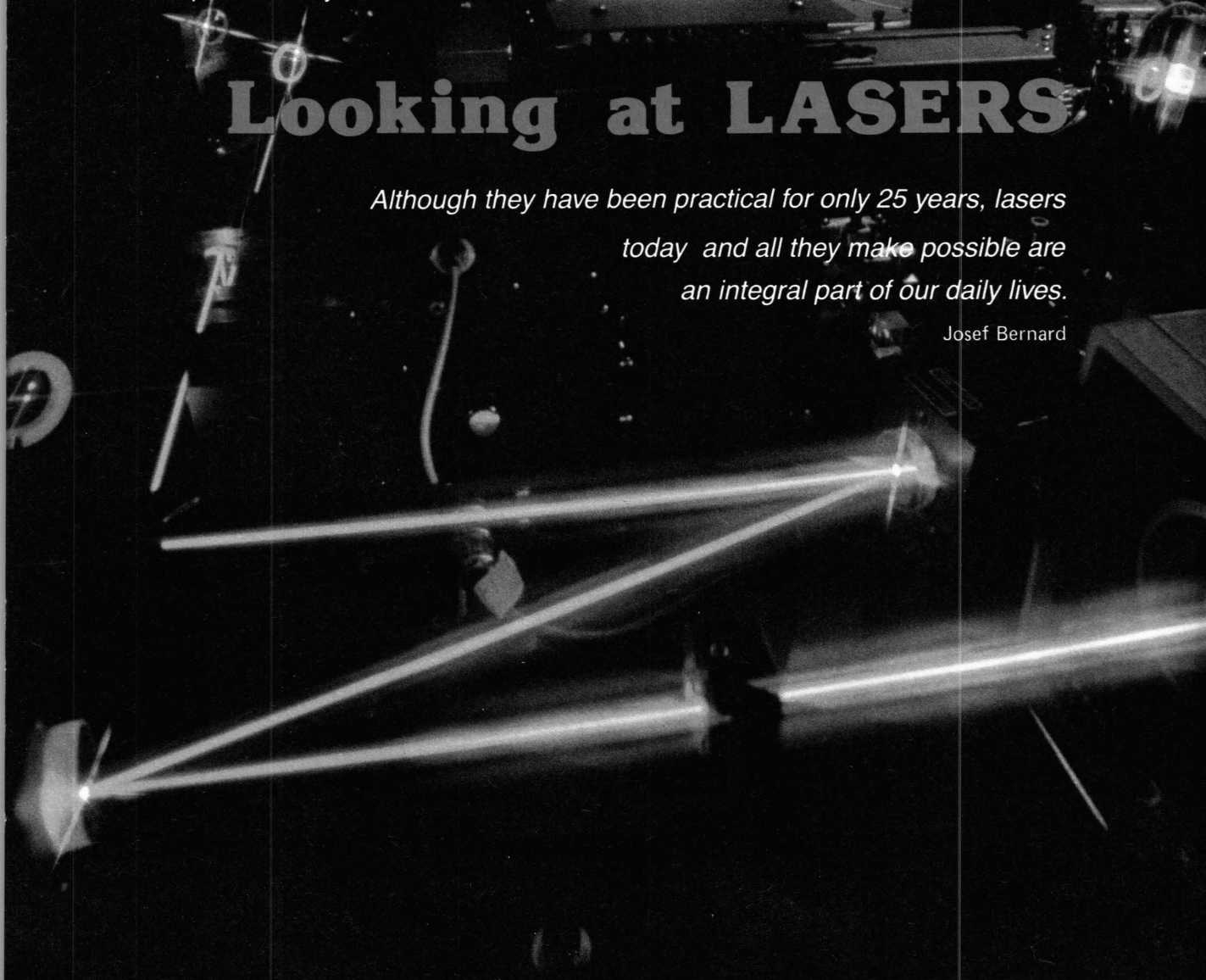
LASERS, LIKE TRANSISTORS AND digital computers, exemplify the way in which what was not too long ago a laboratory curiosity can become an inescapable part of daily life. While you may not yet own a laser-bearing device, you almost certainly use one every day.

Lasers carry communications on fiber-optic cables, play music from CD's, and read prices at supermarket checkout-counters. They perform surgery, help survey our highways, test the components of the airplanes we fly in,

Looking at LASERS

Although they have been practical for only 25 years, lasers today and all they make possible are an integral part of our daily lives.

Josef Bernard



and entertain us at rock concerts. They also make formidable weapons.

What is a laser?

A laser (which stands for *Light Amplification by Stimulated Emission of Radiation*) is a source of intense light that has several unusual and useful properties. The light is monochromatic, which means that it is a single, very pure, color whose frequency can be measured and used as a precision standard in and out of the laboratory. Laser light is coherent—all the waves of a beam are in phase. Unlike natural and most artificial light sources, whose emissions are incoherent, or phased randomly, a laser produces packets of photons all “marching in step”, and possessing a great deal of energy. Finally, because of the way they are generated the rays of light produced by a laser are all parallel to one another, or very nearly so. A pencil-thin beam of laser light aimed at the moon will spread out to a diameter of only 1½ miles. That may not sound impressive—until you consider that the light would travel a distance of about 250,000 miles before diverging that far.

How lasers work

The principle behind the laser is called stimulated emission. That term refers to the fact that an atom, when in an excited state, can be made to emit a photon when bombarded by other photons or by another high-energy source such as a source of electrons.

That needs a little explaining. Figure 1-a shows a simple hydrogen atom—just a proton and an electron. The “height” of the orbit of the electron depends on the amount of energy that the electron is carrying and is very strictly defined by nature. Further, for an electron to be forced to jump from one orbit to another requires a precise amount of energy, or quantum, to be added to it.

In its lowest orbit, no energy has been added and the electron is in its normal, or ground, state. When just the right amount of energy is applied, the electron will jump to a higher orbit, and it is said to be excited (Fig. 1-b). That excited state, though, is unstable, so the electron quickly returns to its ground state. When it does that it gives up the energy that was applied to it, in the form of a photon, or particle of light. (Light is electromagnetic radiation but it frequently behaves like a particle. For that reason, a photon is sometimes called a “waveicle.”) That spontaneous emission of photons is where laser light begins.

If you get enough excited atoms together, spontaneously emitting photons as they undergo the transition from the excited to the ground state, an interesting thing happens. When one of those photons strikes an already excited atom, it changes state and emits its own photon (Fig. 1-c).

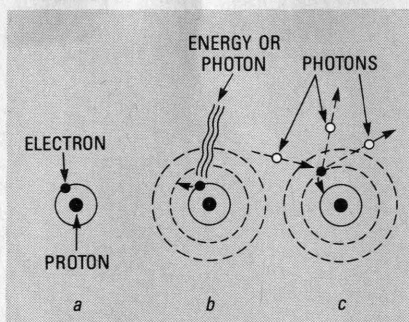


FIG. 1—A SIMPLE HYDROGEN ATOM is shown in a. If a precise amount of energy, or quantum, is added to the atom (b), the electron makes a transition from its normal ground state to a less-stable excited state and jumps to the next higher orbit. If a photon strikes the electron while it is in that excited state, the electron will return to the ground state and emit a photon of its own (c).

Thus, there are then two photons where previously there was one. Those two photons can go on to strike other excited atoms and generate still more photons—a process very much like what happens in a nuclear-fission reaction. That process of photons generating other photons from excited atoms is known as stimulated emission. A photon of a particular wavelength gives rise only to photons of the same wavelength. Thus, laser light is monochromatic; it contains only a single color.

The trick, of course, is to get together in one place a large number of excited atoms of the same kind, so that the stimulated emission of photons can take place. That is done by pumping the atoms in a material up to an excited state by bombarding them with intense light or with some other source of energy such as a beam of electrons. With a ruby laser—the type used by Theodore Maiman on July 7, 1960 when he demonstrated the first laser—as an example, let's examine the lasing process.

A simple diagram of a ruby laser appears in Fig. 2. The heart of that device is a rod of synthetic ruby whose ends are finely ground and polished so they are optically flat and are exactly parallel to one another. Both ends of the rod are

silvered to reflect light back into it, but the reflecting surface at one end is not a perfect reflector—it allows perhaps ten percent of the light generated within the rod to escape. That light is the laser beam.

The rod is surrounded by a spirally-wrapped xenon flash tube similar to those used in electronic flash-units. The light produced by that tube will excite the atoms in the ruby rod. Because of that, that type of laser is known as an optically pumped laser. (As we shall see, there are other types of excitation commonly used.) The cooling equipment is present to remove the heat generated by the lasing device. Lasers are extremely inefficient—only one or two percent of the power they consume is transformed into usable laser light; the rest is given off as ordinary light and lots of heat. That isn't really too bad though—an ordinary incandescent bulb is only about two percent efficient, and the light it produces can't begin to compare with that from a laser.

When the flash tube discharges, the photons it emits enter the ruby rod through its sides and excite the material's chromium atoms, which absorb green and blue light. (Those atoms are what give the ruby its reddish color; you may remember from physics that whatever light a material doesn't reflect or transmit, it absorbs.) When those excited atoms decay from their excited state they give off photons, which trigger other excited atoms to release photons, and so on. The whole process in a ruby laser takes place in about 300 microseconds, and an intense burst of ruby-red light is produced.

We now have lots of light, but we still don't have a laser. That's where the reflective end surfaces of the rod come in. Most of the red light generated within the rod escapes through the sides, but some of it is reflected back into the rod, and that gives rise to the stimulated emission of more red light (hence the “amplification” in the word “laser”).

A portion of the light is not reflected, however, but escapes from the rod through the end that is only partially silvered. That

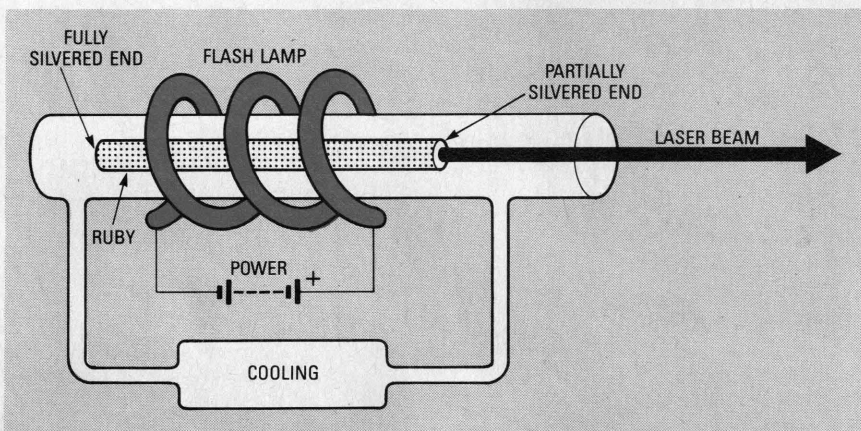


FIG. 2—IN THIS OPTICALLY-PUMPED LASER the light produced by a xenon flash lamp is used to excite the atoms in a ruby rod.

is the laser beam. It is monochromatic because the photons that trigger stimulated emission give rise only to photons like themselves. It is also coherent—all the light waves are in phase. That, too, is a result of the process of stimulated emission; the phase of the photons generated is identical to that of the stimulating photon. The rest of the light reflected by the rod's mirrored ends bounces back to interact with more chromium atoms and produce more photons.

Figures 3-a and 3-b, respectively, represent coherent and incoherent (random phase) light. As is the case with any wave phenomenon—we're now considering photons as waves rather than as particles—out-of-phase waves tend to cancel each other. Because all the waves of a laser beam are in phase, it is much more intense and powerful than a beam of ordinary incoherent light.

Finally, all of the photons in a laser beam travel parallel to one another. That is the result of the orientation of the reflecting surfaces at the ends of the lasing element. The beam of even an inexpensive laser has a divergence of only about one-twentieth of a degree, which means that the energy it carries is not diffused appreciably over distance.

There are many, many types of lasers, and their characteristics and modes of operation tend to overlap. The following examples are just a small cross section of what has been developed in the past 25 years.

Crystal lasers

This is the category to which the original ruby laser belongs. Those lasers are optically pumped and have a relatively low-power output, in the milliwatt range. The most common type is the neodymium-YAG (for Yttrium-Aluminum-Garnet) laser, which emits light in the near-infrared. YAG lasers can be operated continuously because the material from which they're made conducts heat, which would otherwise destroy the laser rod, relatively well.

Another member of the crystal-laser family is the neodymium-glass laser. It is less expensive to produce than the YAG type (glass is cheaper than garnet, even the synthetic kind), but it must be pulsed, or operated on a one-shot basis. It cannot sustain continuous operation because of glass's poor heat conductivity.

Gas lasers

There are more gas lasers than there are any other type. Over 5000 types of laser activity in gases are known. Gas lasers are not usually optically pumped, but are energized by passing an electric current at a potential of several thousand volts through the gas, which is contained in a tube with polished and silvered faces similar to the ends of the ruby rod described earlier. As

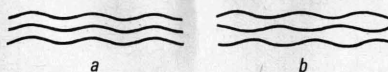


FIG. 3—IN LASER LIGHT, all of the waves are in phase; thus they are said to be coherent (a). In normal light, the waves have no phase relationship (b).

the current flows through the gas, the electrons transfer some of their energy to it, bringing it to a state where the stimulated emission of photons can occur. Because of the way they're constructed, gas lasers can be cooled more efficiently than crystal types, and lend themselves better to continuous operation.

The most widely used gas laser is the helium-argon laser, which can be built for a modest sum by almost any experimenter. It is able to produce no more than 50 milliwatts, but its tight beam of red light, about a millimeter across, makes it ideal for laboratory and experimental use.

Argon and krypton lasers can produce a wide range of colors, but are still relatively low in power. It is not feasible, for example, to construct an argon laser more powerful than 100 watts. Argon lasers with their green light are frequently used in medical applications.

The infrared carbon-dioxide laser is more of a heavyweight. It can have an output as high as several hundred kilowatts. Moderate-sized lasers of that sort are widely used in industry.

Liquid lasers

Organic dyes dissolved in organic compounds such as alcohol can be made to lase, too. Organic lasers are unusual in that one laser can produce a wide range of colors. That spectrum can be optically tuned, and a very precise selection of light of a single color can be made. That capability makes the dye laser a very valuable laboratory tool.

Semiconductor lasers

Semiconductor lasers (Fig. 4) are members of the LED family. They differ from ordinary LED's in that they consume considerably more current and the edges of the semiconductor die are polished to form interior reflecting surfaces. Because of their extremely small size—about as big as a grain of salt—and the difficulty of removing the heat they generate, those lasers do not have a very high output. Still, there are many applications to which they are well suited, among them fiber-optic communications and compact-disc players.

Laser applications

In the 25 years since they came into existence, lasers have proven themselves invaluable in a diverse range of fields. Here are a few of them:

Industry: The high temperatures produced by focused laser beams make them excellent tools for welding, cutting, and

drilling. A pinpoint of coherent light can cut or bore much more cleanly than its mechanical equivalent, with much less waste. (An informal, and entirely unofficial, system for rating the strength of lasers measures their power in terms of "Gillettes"—the number of razor blades that a beam of laser light can successfully punch through.)

Photographs taken by laser light can be used to determine stress regions and faults

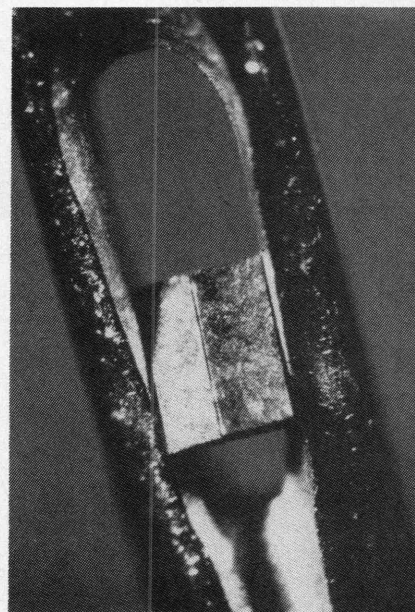


FIG. 4—THIS SEMICONDUCTOR LASER is so tiny that it can fit in the eye of an ordinary sewing needle. Photo courtesy of ATT Bell Laboratories.

in materials, simplifying and improving quality control procedures in critical applications. Lasers are also used in industry for non-contact monitoring of a wide variety of systems. See Fig. 5.

Medicine: Lasers find applications in numerous areas of medicine, among them dermatology, gynecology, and many areas of surgery. The finely focused beam of a laser can operate in areas (such as the inside of the eye) inaccessible to the traditional scalpel.

Science: Lasers have helped scientists both to refine existing knowledge and to learn more about our universe. Using lasers, it has been possible to determine



FIG. 5—LASERS HAVE many applications in industry. Here's a device that uses a laser to monitor the performance of an ultrasonic wire bonder, which is used in electronics production.

the speed of light (186,282.398 miles/second; 299,792.458 kilometers/second) with an accuracy hitherto unknown, and other units of measures have also benefited. A laser beam follows what must be the world's straightest line, a boon for surveyors and the like. Lasers in the laboratory have also allowed the development of new techniques to perform tasks that were previously impossible. Nuclear fusion reactions making possible the generation of enormous quantities of inexpensive electricity from plain seawater will probably be initiated and sustained by lasers.

Communications: Right now fiber-optic communications links using semiconductor lasers are in limited use, but their potential for carrying vast quantities of information makes it certain that as new installations are made, they will become much more common. In space, where laser light cannot be attenuated by air, it may carry communications and data from satellite to satellite, or even to earth. Lasers also are the heart, of course, of the laser printers; those devices, with their high-quality outputs, are now becoming popular in computer circles.

Entertainment: Laser-light shows are popular at rock concerts, and lasers are also used to record and read the information contained on CD's and most videodiscs. Holography, practical only with laser light, makes possible 3-D photography without a camera or special viewing device, and has given birth to a new art form. One day we may enjoy holographic movies, although holographic television at this point seems rather farfetched because of the limited resolution of even the most sophisticated video systems. The applications of holography, of course, are not limited to the world of entertainment. Holographic techniques are also used in devices like scanners for UPC (Universal Product Code) readers in stores, and in the restoration of artwork.

War: Like dynamite, lasers can be put to both peaceful and destructive uses. Currently in the headlines is the "Star Wars" technology that will take the science of war into the peace of space. Lasers are also used in the navigation systems of missiles and in targeting devices.

New uses for the unique qualities of laser light are constantly being conceived. Among some of the more unusual and esoteric areas being explored are dental holography, gene manipulation, acupuncture, laser-based optical computers, and the use of lasers to transmit power from solar-energy-gathering satellites. Future applications of the laser may only be limited by the scope of human imagination.

It doesn't take much to see that the invention of the laser is one of the most significant things to come out of the laboratory in this century.

R-E

BUILD THIS

HELIUM — NEON LASER

*Build this simple helium-neon laser and start
having fun with photons!*

ROBERT GROSSBLATT and ROBERT IANNINI

BACK IN THE HEYDAY OF SCIENCE FICTION's era of purple prose, tales of bug-eyed monsters, death rays, and the like filled many a pulp magazine. Of course, we knew then that it was all just fantasy; you could no more have a "death ray" than you could travel faster than sound or put a man on the moon.

While those bug-eyed monsters (or BEM's, for short) have yet to pay us a visit (to the best of our knowledge), much of yesterday's science fiction is today's science fact. We even have a death ray, of sorts. Of course, we are referring to the laser, which can be a powerful weapon in the hands of those who wish to use it as such.

But the laser is also a great tool for science and industry. In just 25 years the laser has gone from far-fetched notion, to scientific reality, to common noun. Hardly a day goes by where some part of our lives is not affected by lasers. Today, the laser has joined the transistor as a hallmark of modern electronics.

What's a laser?

The word *laser* is an acronym for *Light Amplification by Stimulated Emission of Radiation*. But for most of us, that provides a poor explanation of what a laser is and how it works. To find a better explanation, we have to leave electronics for a while, drop into the world of physics, and talk a little bit about the nature of

light. You can't understand laser light until you have some familiarity with the properties of light in general.

There are three ways in which laser light differs from ordinary light, and each of those differences contributes to the special characteristics of a laser. Let's begin by looking at some of the characteristics of ordinary light.

Ordinary light has a relatively wide bandwidth. That means that a spectrographic analysis would reveal that regular light is made up of many different wavelengths. Just about everybody has seen, or done, the experiment in which a beam of white light is directed through a prism and split into different colors. The ordinary light we see as white, therefore, is actually made up of different color elements—it's *polychromatic*. Figure 1 shows the composition of visible light, and the relative sensitivity of the human eye to various wavelengths.

Ordinary light is also *temporally incoherent*. By that we mean that the various components of the light do not share any time relationship; they are all randomly out-of-phase with respect to each other. Thus, if you were able to look at the waveform of a beam of ordinary light, you would see something that looks like Fig. 2. The irregularity and random appearance of that waveform is caused by the presence of waveforms of differing frequencies in the light, and the ways in





which those waveforms interact with each others. Because of that interaction, at some points the waveforms all add (constructive interference); at some points they all subtract from one another (destructive interference), and at most points a combination of the two effects occurs. The result is an irregular, random-appearing waveform as shown.

Finally, ordinary light is also *spatially incoherent*. If you were to analyze the waveform of Fig. 2 over a period of time you would see that it is constantly changing. That's because the component waveforms are constantly changing their positions with respect to each other, causing the interference effects to vary.

The best way to put the differences between ordinary and laser light in perspective is to compare light to sound. Ordinary light, because of all the things we just talked about, can best be compared to noise. The waveforms at any moment in time are not only randomly spaced, but there's an unpredictable mix of frequencies as well.

Now, if regular light is like noise, then laser light can only be thought of as the purest sound imaginable. For openers, laser light is highly monochromatic—a spectrographic analysis would show that it is composed of light of only one wavelength. And where regular light is temporally incoherent, a laser is temporally coherent—all of the light waveforms are in phase with each other. That is one of the reasons why a laser puts out light of such pure color. Being monochromatic helps, of course, but being temporally coherent as well means that there's almost a complete absence of what would be called distortion in a sound wave.

As you might have already guessed, laser light is also spatially coherent. If you looked at the waveforms over a period of time, there would be absolutely no shifting or movement. Considering the absence of interference effects, that is exactly what you would expect to happen.

Taken together, all of those factors are what make laser light so intense, and so directional.

Since laser light is just about interference free, there's almost no scattering of the light. The beam divergence is very small—in the milliradian range. A laser beam is really a beam of light! Being coherent also means that there's a much smaller loss in energy over distance than there is with regular light. Obviously, since laser light is so different from regular light, it can't be produced the same way. And in order for us to understand how it's produced, let's see how regular light is produced.

Electromagnetic waves in general, and light in particular, is produced when an atom gives off energy. Now, an atom either takes on energy (absorption), or gives off energy (emission), by having its electrons move from one energy level to another. Once energy has been supplied to the system, and absorbed by the atom, emission can occur in one of two ways—it can happen spontaneously, or it can be stimulated.

Spontaneous emission is the result of natural atomic decay. The electrons randomly drop in energy level and produce the kind of waveforms shown in Fig. 2. When you power up a light bulb, for example, the atoms in the filament absorb energy and release it as a combination of heat and ordinary, incoherent light.

Stimulated emission is a completely different process. The idea is to keep the atoms from releasing their absorbed energy in a random manner. In order to do that, you have to create a state of affairs called a "population inversion." In simple terms, that means that there have to be

more atoms present that have absorbed energy than there are atoms that have already released energy. That is usually brought about by forcing enough energy into the system to create at least three atomic-energy levels. The most excited atoms will emit energy and drop down to the next lowest energy level. That means there will be more atoms at the intermediate level than there are at the lowest level—that's the population inversion we mentioned.

As photons of energy are released, they will cause other atoms to emit energy in the form of photons. Those photons will trigger the production of other photons. And if the emission is bounced back and forth between two mirrors the production of photons will continue to build in phase and the result will be, you guessed it, a beam of laser light with a waveform that looks like that shown in Fig. 3.

Making a laser

Now, understanding the basic theory and putting it into practice are, as we all know, two completely different things. Creating the population inversion you need to produce a laser beam is really an iffy, ticklish business. Everything has to be just so or nothing will happen. The mirrors have to be of a certain type to produce the in-phase coherent energy needed for a laser. And enough energy of the right type has to be forced into the system to make the whole thing work.

The kind of energy you have to pump into the system depends on the type of material you're trying to make lase. Semiconductor and gas lasers are pumped up with electrical energy while crystalline lasers, such as those made from ruby rods or YAG (Yttrium-Aluminum-Garnet) are usually pumped up optically with xenon flash tubes or arc lamps.

The laser we're building here is a gas

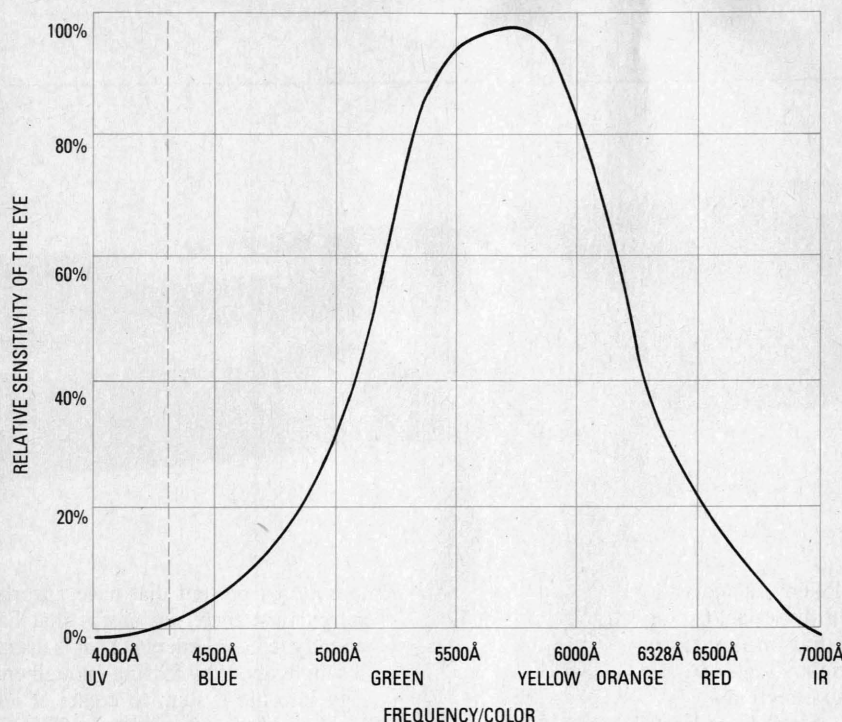


FIG. 1—THE VISIBLE SPECTRUM, and how the human eye responds to it. The wavelength of the light emitted by our helium-neon laser is 6328 Angstroms.

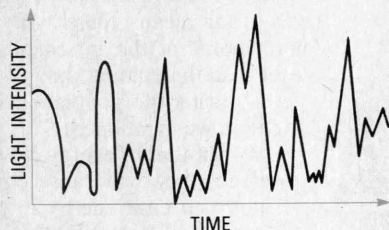


FIG. 2—THIS RANDOM-APPEARING waveform is that of ordinary light. The waveform is made up of all of the various frequencies that make up such light.

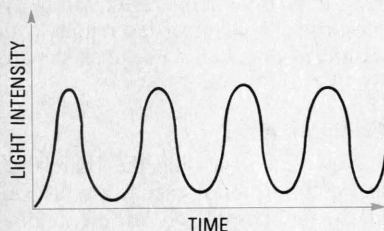


FIG. 3—LASER LIGHT is made up of light of just one frequency. It is the purest type of light possible.

laser—more specifically a helium-neon laser. The frequency of the light is 6328 Angstroms and the laser puts out about 1 milliwatt with a beam divergence of 1.3 milliradians. Now, 1 milliwatt may not sound like a lot of to you, but that's because you're still thinking in terms of regular light. Remember that the laser produces a highly directional beam of coherent, monochromatic light. The laser we're talking about here generates a beam that can be spotted on a wall more than two miles away!

Helium-neon lasers are extremely inefficient and, in order to make them work,

the mechanical setup of the laser tube has to be just about perfect. It has to be properly sealed and contain the correct gas mixture. Also, the mirrors have to be perfectly aligned dielectric ones so enough reflection takes place at the proper frequency to cause the device to lase. Those mirrors must be highly reflective, within a couple of decimal places of 100%; by contrast, the silver mirrors we use every day have a reflectance factor of only 95%.

Making a helium-neon laser tube is a project that is beyond the means of most of us as it requires a fair amount of skill and equipment. Among other things, you need to have the skills and equipment required to create a precise mixture of gases, and you need to be adept at glass blowing. All of that is not impossible, of course, but in most cases it's a task that is best left to someone else; we recommend that you purchase rather than build a tube. (One source for laser tubes is mentioned in the Parts List.)

Once you have a working laser tube, actually making it produce a beam is surprisingly simple. The only electronic assembly needed is a power supply that will deliver the right voltage to make the tube fire. Figure 4 is the schematic of a power supply that can be used to trigger the laser. If it looks familiar, that's because its front end is essentially the same one used in the construction of the infrared viewer that appeared in the August 1985 issue of *Radio-Electronics*.

The power supply is a switcher with Q1, Q2, and their related components forming an oscillator that switches a square-wave through the primary windings of T1,

a high voltage step-up transformer. That part of the circuit takes the battery voltage and produces about 400 volts AC at the secondary of T1. Diodes D3–D6 and capacitors C2–C5 form a voltage multiplier that takes the 400 volts from T1 and boosts it to the 1600-volts DC needed to ignite the laser tube.

The high-voltage pulse needed to ignite the tube comes from an 800-volt tap on the voltage multiplier. Resistors R3 and R4 divide that voltage to provide the 400 volts needed to charge up C9, the dump capacitor. When the SCR fires, the charge on C9 is dumped into the primary of the trigger coil, T2. Capacitor C11 charges up and, since it's in parallel with the laser tube, when the voltage builds enough to excite the gas, ignition takes place and current flows through the tube. That causes a voltage drop across R10, which turns on Q4 and turns off Q3.

As soon as the laser tube ignites, therefore, the ignition circuitry is turned off. That saves battery power because the laser tube can sustain firing at a lower voltage. The relaxation oscillator made up of Q3 and Q4, and their related components is only needed to control the firing of the SCR. Once the tube starts to lase, the voltage drop across R10 keeps the ignition circuitry turned off. If the tube stops lasing, the R9–R10 junction will drop to ground again and Q4 will turn off and unclamp Q3. The SCR will start firing again and, we hope, re-ignite the tube.

Construction

Before we actually start building the circuit, there's one very important thing you *must* keep in mind:

CAUTION! The power supply can produce as much as 10,000 volts at about 5 milliamps. That is enough juice to do a lot of damage. If you're not careful you can give yourself a severe shock. Remember that the capacitors take a while to discharge completely. You can get a real jolt even if the circuit has been turned off for five or ten minutes. Treat the circuit with respect and make sure to discharge the capacitors if you want to do some work on the circuit.

Now that that's out of the way, you can build the power supply on perfboard or use the PC board that's provided in our PC Service section, elsewhere in this magazine. If you use perfboard, remember to keep the leads as short as possible because there's a lot of high-frequency AC running around part of the circuit. Whichever method you use, make sure to keep any metal objects and your fingers away from the output section located around T2 and R11. Those are the points of the circuit where the highest voltages can be found. One short second of carelessness on your part and you're going to get zapped. If you're lucky, all it will do is hurt a lot.

The only other components in the cir-

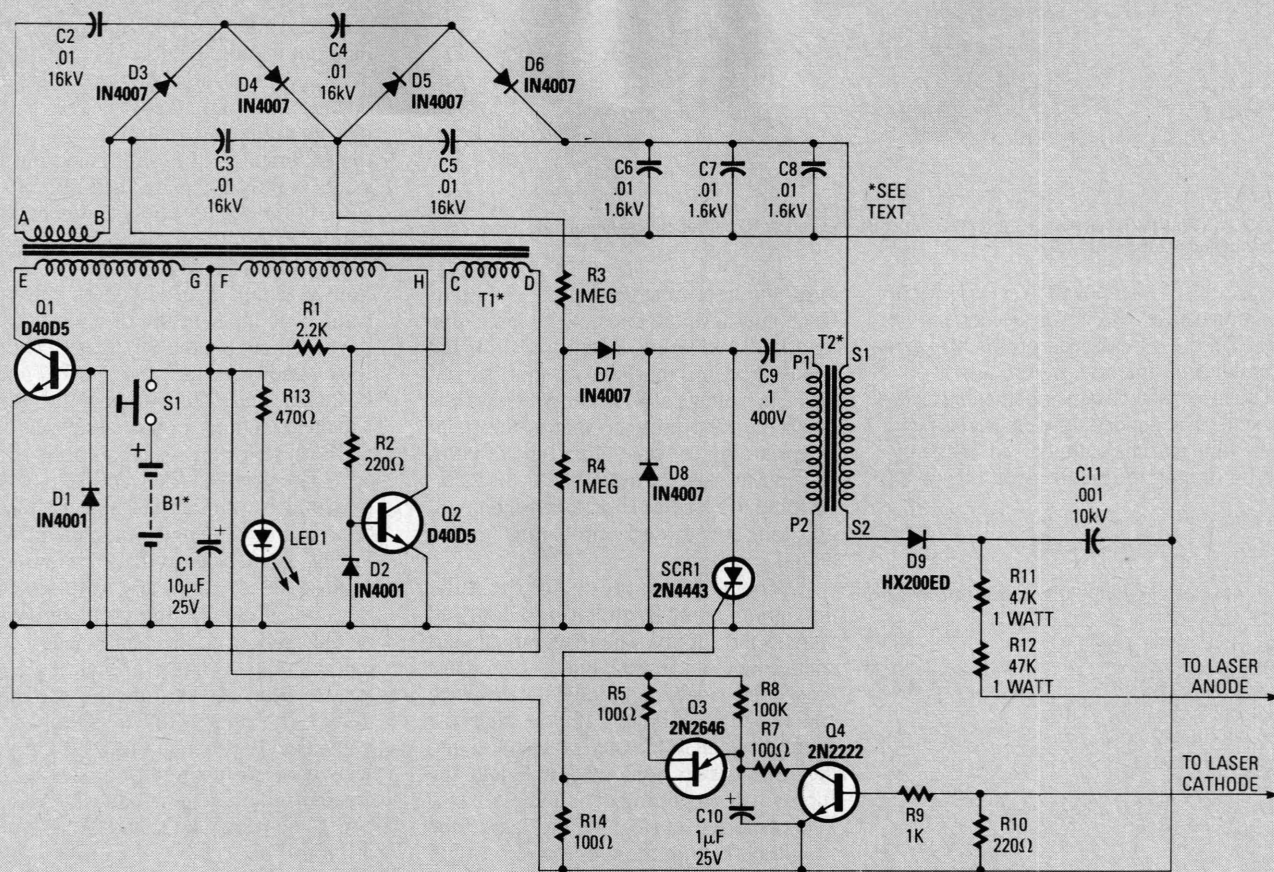


FIG. 4—THIS POWER SUPPLY is all you need to drive a laser tube like the one available from the supplier mentioned in the Parts List.

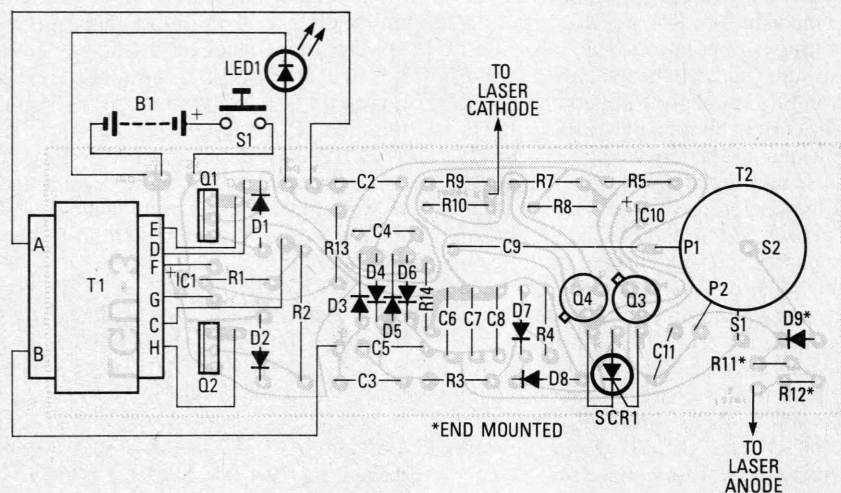


FIG. 5—IF YOU CHOOSE to use the PC board provided in our PC Service section, use this parts-placement diagram.

cuit that require special attention are the switching transistors, Q1 and Q2. The maximum current draw from the batteries is about 750 mA, so those transistors will be handling a lot of juice and getting hot. The PC-board layout shown in Fig. 5 is designed so that the transistors can be placed in such a way that their tabs can be stuck against the laminations of T1. If you are using perforated construction board, be sure that your layout allows for that, too. Use some heat-sink compound to get good thermal contact, and using small

heat sinks wouldn't be a bad idea.

After you've identified the components and found their position on the board, solder them in using a minimum of solder. Once you've done that, use some high-voltage putty, paraffin, or varnish to cover the traces (or wires if you're using perf-board) that connect to all the components on the secondary side of T2 and the laser tube. That part of the circuit has the highest voltages and it's likely that arcing will take place if all the bare metal isn't covered. You may find it necessary to use the

same material on the component side of the board as well.

When you finish the board, check for bridges, opens, bad solder joints, and so on. If everything seems OK, you're ready to test the power supply. Take the two leads that normally would go to the laser tube and tape them down so that they're 1/8-inch apart. Connect 10 volts to the power supply. You should see arcing across the laser-tube leads at a rate of about once a second or so; the circuit should be drawing approximately 250 mA. If the spark becomes continuous, the current draw should jump to about 750 mA—the full operating current of the laser tube. If you measure the voltage across the output of the supply, you should see an open circuit voltage of about 2500. Once the laser tube is connected, the voltage will be in the neighborhood of 1500.

If you've gotten this far without any brain damage, you're ready to connect the tube to the supply.

CAUTION! The laser tube is an expensive, delicate piece of equipment. In order to connect it to the circuit you'll be soldering leads to the metal collars at either end of the tube. Use a minimum of solder and apply heat for a minimum amount of time. Don't ever forget that the tube has a high vacuum inside and you can damage more than the tube if you destroy the integrity of

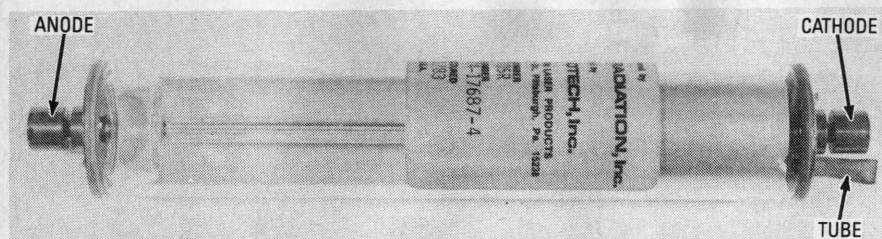


FIG. 6—A HELIUM-NEON laser tube. The cathode end can be identified by the small tube used to fill the laser tube with gas.

the seal. Use a low-power iron and a lot of common sense when you solder to the tube. Tin the wires ahead of time to keep the soldering time to a minimum.

The laser tube has an anode and a cathode end. The anode is the clear glass end of the tube and the cathode can be identified by finding the small metal tube used

to fill the laser tube with gas. See Fig. 6. Once you have the ends of the laser tube identified, solder the lead from R12 to the anode, keeping that lead as short as you can. The voltage at R12 is about 1100 volts when the tube fires, so that is the point where arcing or some other type of parasitic power loss is most likely to occur. Keeping the lead as short as possible will go a long way toward eliminating any potential problems.

Connect the cathode of the tube through an ammeter to R10 and apply 10 volts to the supply. The tube should start sputtering as it tries to ignite. As you raise the voltage *slowly*, ignition will take place and the tube will fire. If you raise the supply voltage to 12 volts, you should find the tube drawing about 5 mA and the supply will be putting out about 750 mA. The operating current of the laser tube should be kept in the range of 4.5 to 5.5 mA. Less than that and the tube won't operate reliably; more than that and you'll cut down its operating lifetime. You can trim the values of R11 and R12 to get the operating current into the safe range. Just make sure you keep the total resistance of R11 and R12 at about 100,000 ohms.

The current through the tube is going to vary a lot as the battery voltage changes. If you use nickel-cadmium cells, remember that the operating voltage is going to be 1.2 in each cell for most of the lifetime of each charge. Freshly charged nickel-cadmium cells, however, can have a voltage as high as 1.5. Admittedly that doesn't last very long, but we'll mention it because adjusting the laser tube's operating current with freshly charged batteries could cause you to choose artificially high values for R11 and R12. So before you start trimming the resistors, make sure the batteries are at 1.2 volts per cell.

Since the operating current is tied to the supply voltage, it's natural to think about voltage regulation. Well, there is nothing wrong with regulating the input voltage, but there are a few things to keep in mind. If you use 12 nickel-cadmium cells, you'll have a supply voltage of 14.4 and you'll be drawing as much as 750 mA from that supply. A three-terminal regulator like the 7812 would seem to be an ideal choice, but asking it to supply 750 mA is really asking a lot unless you use a 7812C, which can handle 1 amp with a good heat sink. An LM317 can be set up to put out 12 volts and it can supply 750 mA.

The biggest problem with using an IC voltage-regulator is the voltage loss that's inherent in those devices. In order to supply 12 volts, a regulator needs an input voltage of about 14.5 volts. Now that's just about the maximum you can get from the batteries. And if your particular tube wants a little bit more than 12 volts, or some of the power-supply components are a little bit lossy, you're in a lot of trouble.

So, you ask, what's the bottom line. Well, after all's said and done, unless you want to do an awful lot of circuit design, the best thing to do is let the power supply look directly at the batteries. It's not the best solution in the world, but it's probably the best thing in this situation.

The case for the laser can be as simple or as fancy as you like. Perhaps the simplest and most functional approach would be to use some lengths of standard PVC tubing. But if you do that, or completely enclose the circuit in any way, you could run into an overheating problem because of the amount of heat produced by the power supply. Because of that, it's a good idea to limit the on-time to less than a minute; keeping it under 30 seconds is even better. Further, giving the supply a 5-second or so rest between uses will increase its lifetime tremendously. Also, the better you heatsink Q1 and Q2, the better off you'll be.

Having fun

The output of the laser tube is about 1 milliwatt (or 0.4 milliwatt if the lower-powered tube offered by the supplier mentioned in the Parts List is used) and, at that power, it can't do any damage. If you had thoughts of burning your way through steel, forget it. Lasers that can do that are worlds away from the one we're building. However, that doesn't mean you can treat the light from this laser with no respect whatsoever.

CAUTION! Even a 1-milliwatt laser can be hazardous if you look directly at the beam. While we assume that anyone considering building a laser would know enough about those devices to never, never even consider doing something so foolhardy, the very nature of laser might make it very easy for accidents to happen. The beam is highly directional and very intense; to compound matters, the reflected beam is just as dangerous as the emitted beam. It's a simple matter to have the beam bounce off some shiny object and reflect back to you. You can wear safety glasses, but even if you do, be careful where and how you use the laser.

While you can use this laser, which throws an intense red beam, for such things as target spotting, perhaps its greatest use is as an introduction to the world of lasers in general. Watching the tube fire is truly fascinating and the more you experiment with it, the more you'll learn.

R-E

PARTS LIST

All resistors ¼ watt, 10% unless noted

- R1—2200 ohms
- R2—220 ohms, 1 watt
- R3, R4—1 megohm
- R5, R7—100 ohms
- R6—not used
- R8—100,000 ohms
- R9—1000 ohms
- R10—220 ohms
- R11, R12—47,000 ohms, 1 watt
- R13—470 ohms

Capacitors

- C1—10 μ F, 25 volts, electrolytic
- C2—C8—0.01 μ F, 1.6 kV, ceramic disc
- C9—0.1 μ F, 400 volts, paper dielectric
- C10—1 μ F, 50 volts, electrolytic
- C11—0.001 μ F, 10 kV, ceramic

Semiconductors

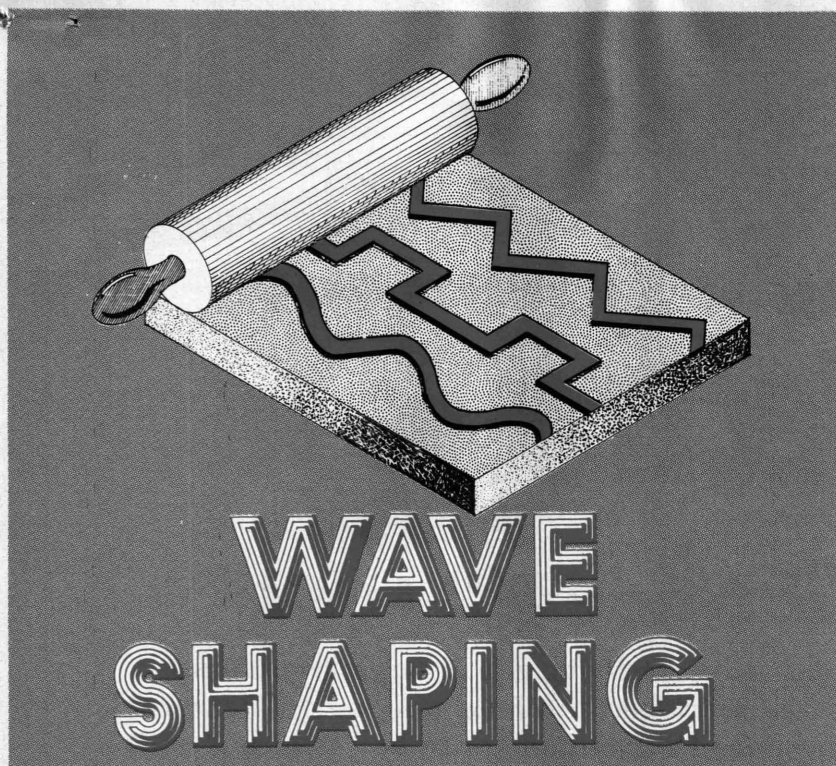
- D1, D2—1N4001
- D3—D8—1N4007
- D9—HX200ED, 20 kV diode
- LED1—Red LED
- Q1, Q2—D40D5, NPN power transistor
- Q3—2N2646—UJT transistor
- Q4—PN2222 NPN transistor
- SCR1—2N4443 SCR

Other components

- T1—12 to 400 volts, 10 kHz switching transformer
- T2—10-kV trigger transformer, 400-volt primary
- B1—14.4 volts, 12 nickel-cadmium cells, or equivalent
- S1—SPST switch, momentary pushbutton, normally open

Miscellaneous: PC board, helium-neon laser tube, PVC tubing for case, battery holders, wire, solder, etc.

Note: The following are available from Information Unlimited, PO Box 716, Amherst, NH 03031: PC board, \$4.50; switching transformer (T1), \$14.50; trigger transformer (T2), \$11.50; 1-milliwatt laser tube, \$149.50; 0.4-milliwatt laser tube, \$99.50; high-voltage diode (D9), \$3.50; high-voltage capacitor (C11), \$3.00.



Learn the basics of waveform generation and shaping with bipolar transistor circuits that you can build and put to work.

RAY MARSTON

THE SUBJECT OF THIS ARTICLE IS waveform generation and shaping as performed by various kinds of multivibrator circuits and special-purpose oscillators. It is a continuation of last month's article on transistorized RC and LC oscillator circuits, and the astable multivibrator. Previous articles in this series have covered the basics of the bipolar junction transistor (BJT) and have presented a general roundup of popular BJT circuits starting with those basic transistor amplifiers: common-collector, common-emitter and common-base.

Multivibrator basics

A transistor multivibrator is a cross-coupled, two-stage switching circuit. Each active transistor stage is regeneratively cross-coupled to its companion; thus, one stage automatically turns on as the other turns off, and conversely.

This cross-coupling can be arranged to give either stable or semistable switching. When stable cross-coupling is desired, the transistor switch locks permanently into the ON or OFF state until it is forced to change state by an external signal.

When the circuit is cross-coupled in a semistable manner, the transistor initially locks into the ON or OFF state, but then automatically becomes "unlocked" again after a delay period determined by the time constant of the cross-coupling components.

Schematics of the four basic transistor multivibrator circuits most commonly used are shown in Figs. 1 to 4. The Fig. 1 circuit is a manually triggered bistable (two stable state) multivibrator. The base-bias of each transistor is obtained from the collector of the other transistor, so that one transistor automatically turns off when the other turns on, and conversely.

The output can be driven low by briefly turning Q1 off with

S1: the circuit locks into this state until Q2 is turned off by S2. At that time the output locks into the high state, and this action can be repeated as long as the circuit is powered.

Figure 2 shows a monostable (one stable state) multivibrator or one-shot pulse generator circuit. Its output is normally low, but switches high for a preset period (determined by the values of C1 and R2) if Q2 is briefly turned off with S1.

Figure 3 shows an astable (no stable states) multivibrator or free-running square-wave generator. The on and off periods of the square wave are determined by the values of R3 and C1 and R2 and C2.

Figure 4 shows a Schmitt trigger or sine-to-square waveform converter. Transistor Q2 switches abruptly from the ON state to the OFF state, or conversely, as the base of transistor Q1 base

risers above or falls below the predetermined trigger-voltage levels.

Several different practical astable multivibrator circuits were discussed in last month's article. This article will examine practical versions of three other multivibrators.

Monostable circuits

The monostable multivibrator circuit in Fig. 2 acts as a triggered pulse generator. Normally transistor Q2 is driven into saturation through R2, so the output (taken from transistor Q2's collector) is low. Transistor Q1, which derives its base-bias from transistor Q2's collector through resistor R4, is cut off under this condition, and its collector is at the full supply voltage.

When a START signal is applied to Q2 by momentarily closing switch S1, Q2 switches off, driving the output high and driving Q1 on through R4. Regenerative switching action is caused by the reopening of S1.

Transistor Q2's base is driven negative by the charge on C1, and as soon as the regenerative response is complete, C1 starts to discharge through R2. Eventually its charge falls so low that Q2 turns on again, thus initiating another regenerative response. Now both transistors revert to their original states, and the output pulse terminates, completing the action of the circuit.

Thus, a positive-going pulse is developed at the output of this circuit each time an input trigger signal is applied by momentarily closing switch S1. The pulse period is determined by the values of R2 and C1. The relationship is:

Pulse period = $\approx 0.7 \times R2 \times C1$
Where the pulse period is in microseconds, C is in microfarads, and R is in kilohms.

The circuit in Fig. 2 can be triggered either manually by closing a momentary switch or by introducing an input trigger signal. That trigger signal can be either a negative pulse applied to the base of Q2, or a positive pulse applied to the base of Q1.

Figure 5-a is a practical schematic for a manually triggered monostable multivibrator. It can be triggered with momentary switch S1 by feeding a positive pulse to Q1's base through R2. Figure 5-b shows the circuit's waveforms.

In Fig. 5, the base-to-emitter junction of Q2 is reverse-biased during the operating cycle by a peak voltage equal to the supply voltage. This means that the maximum supply voltage should be limited to about 9 volts to prevent damage to the transistor. However, a supply voltage greater than the reverse base-emitter breakdown value of Q2 can be applied safely if silicon diode D1 is placed in series with Q2's base, as shown in Fig. 5.

This higher supply voltage provides the same kind of *frequency correction* that was described for the astable multivibrator in last month's article.

The value of timing resistor R3 in the Fig. 5 circuit must be large with respect to R1, but

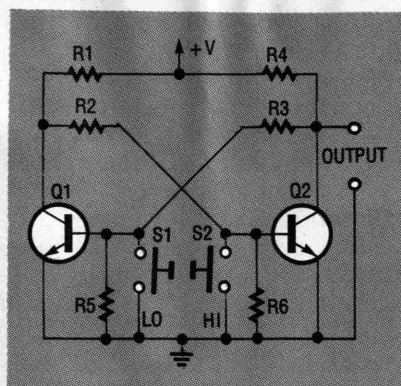


FIG. 1—A BISTABLE MULTIVIBRATOR intended for manual-triggering.

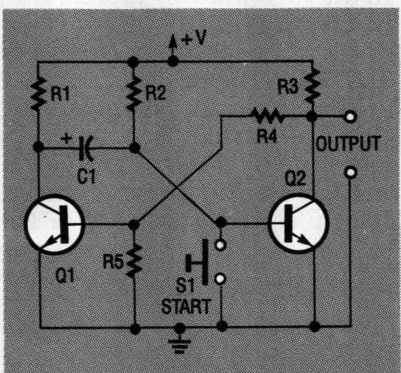


FIG. 2—A MONOSTABLE multivibrator designed for manual triggering.

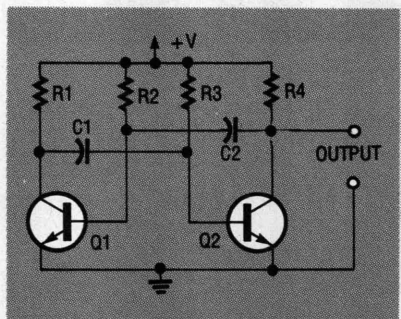


FIG. 3—AN ASTABLE MULTIVIBRATOR or free-running squarewave generator.

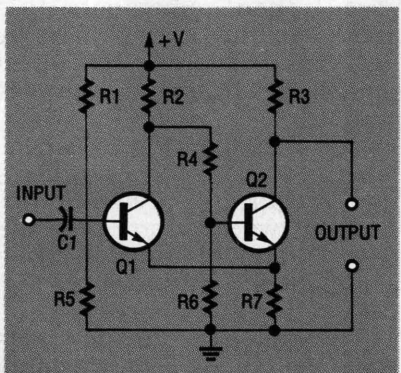


FIG. 4—A SCHMITT TRIGGER circuit is a sinewave-to-square wave converter.

must be less than the product of R5 and the h_{FE} of Q1. The pulse period for Fig. 5 equals 50 milliseconds divided by the value of capacitor C1 in microfarads; it will be 5 seconds with the value of C1 shown.

Long delays

If a Darlington transistor pair is substituted in place of Q2 in Fig. 5, the circuit will be able to provide very long timing periods. That substitution results in a very high effective h_{FE} , and permits the use of large values of R3, as shown in Fig. 6.

The Fig. 6 circuit can be powered from any DC source with an output between +6 and +15 volts to give a pulse output period of about 100 seconds with the values of the resistors and capacitors shown.

Keep in mind that a manually triggered monostable circuit such as those of Figs. 5 and 6 is dependent on the duration of the input trigger signal. The circuits trigger at the moment that a positive-going pulse is applied to the base of Q1 in Fig. 5 or Q3 in Fig. 6. If this pulse is removed before the monostable multivibrator completes its normal timing period, the period will end regeneratively, as previously described.

However, if the trigger signal has not been removed by the time the monostable completes its natural timing period, the timing cycle will end non-regeneratively. This means that the output pulse will have a longer period and falltime than if the trigger signal were removed earlier.

Waveform triggering

Figures 7 and 8 show alternative ways of applying input signal triggering to the monostable pulse generator. In each case, the circuit is triggered by a square-wave input signal with a short rise time. This waveform is differentiated by the differentiation circuit consisting of C1 and R1 to produce a brief trigger pulse.

In the Fig. 7 circuit, the differentiated input signal is rectified by diode D1 to provide a positive trigger pulse on the

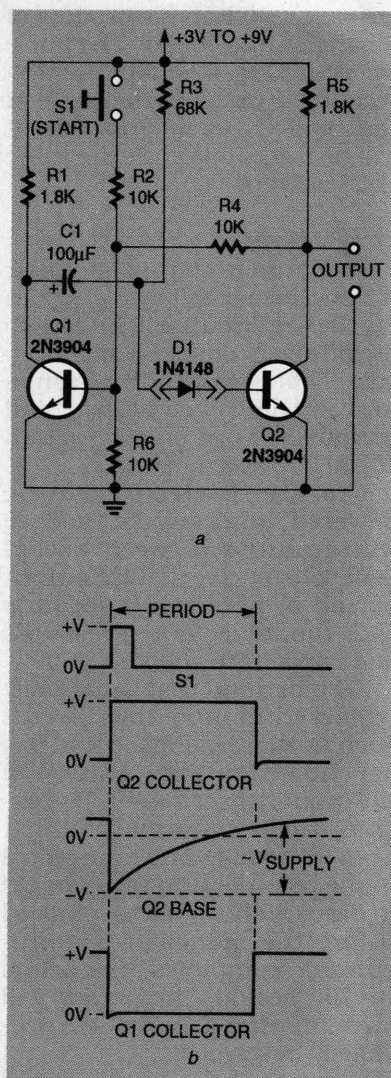


FIG. 5—A MANUALLY-TRIGGERED monostable pulse generator.

base of Q1 each time an external trigger signal is applied. In the Fig. 8 circuit, however, the differentiated signal is fed to the gate of transistor Q1. That change in the circuit makes the trigger signal independent of Q2. Notice that "speed-up" capacitor C3 in Fig. 8 is connected in parallel with feedback resistor R5 to improve the shape of the output pulse.

Both the circuits in Figs. 7 and 8 provide an output pulse period of about 110 microseconds with the values of resistors and capacitors shown. This period can be varied from a fraction of a microsecond to several seconds with a suitable choice of values for capacitor C2 and resistor R4.

The circuits in Figs. 7 and 8 can be triggered by sine or other

non-rectangular waves if they are conditioned by a Schmitt trigger or similar sinewave-to-squarewave converter circuit. (The Schmitt trigger circuit is discussed later in this article.)

Bistable circuits

Figure 9 is practical schematic for the manually-triggered bistable multivibrator shown in Fig. 1 and described earlier. This circuit is also known as a R-S (reset-set) flip-flop and, like a toggle switch, it is also an elementary digital memory. Its output can be SET to the high state by momentarily closing switch S2. (Alternatively a negative pulse can be applied to the base of Q2.)

The circuit then "remembers" this state until it is RESET to the low state by a momentary closing of S1 (or by applying a negative pulse to the base of Q1). The

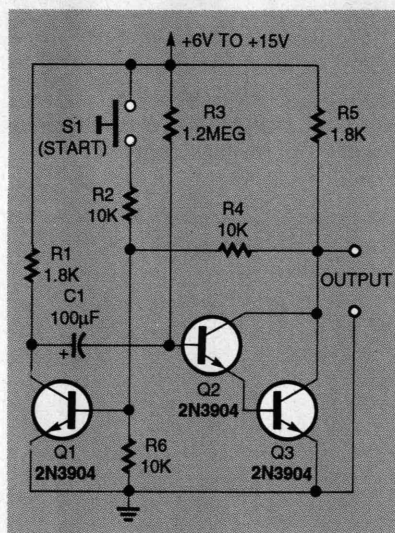


FIG. 6—A LONG-PERIOD (100-SECOND) monostable circuit.

circuit then "remembers" this new state until it is again set by S2. This cycle can be continued indefinitely as long as power is applied.

The circuit in Fig. 9 can be modified to provide a divide-by-two or counting function by including two steering diodes (diodes D1 and D2) and associated components, as shown in Fig. 10.

The Fig. 10 circuit changes state each time a negative-going trigger pulse is applied. If, for example, the input pulses are

derived from a squarewave input signal, the circuit will generate a squarewave output signal at half the input frequency.

The circuit generates a pair of output signals that are 180° out of phase, shown here as Q1 and Q2. The introduction of CMOS IC versions of the bistable counter circuit have largely eliminated any need for the construction of these circuits from discrete components.

Schmitt trigger

The last member of the multivibrator family to be discussed here is the Schmitt trigger circuit. It is a voltage-sensitive switching circuit that changes its output state when the input signal exceeds or falls below preset upper and lower threshold levels. Figure 11 shows how the Schmitt trigger converts sine waves to square waves.

The Schmitt trigger circuit is emitter-coupled and has cross-coupling between the base and collector of transistor Q1, which provides the required regenerative switching. Capacitor C2 speeds up the switching action by shunting R4. The sine-wave input signal is superimposed on a DC voltage. (The voltage is determined by trimmer potentiometer R8 and resistors R1 and R2) that is applied to the base of Q1.

A practical Schmitt trigger needs a sinewave input signal with an amplitude of at least 0.5

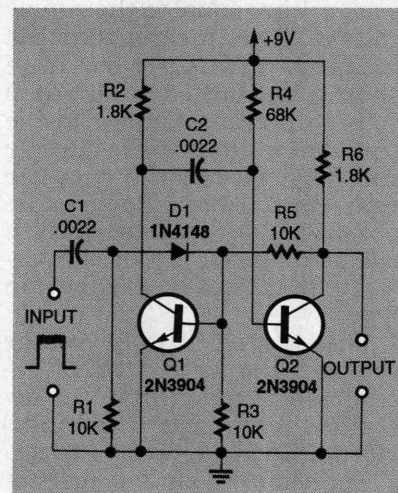


FIG. 7—A WAVEFORM-TRIGGERED monostable circuit.

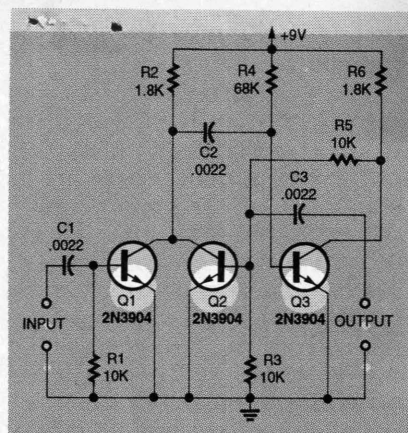


FIG. 8—A MONOSTABLE CIRCUIT with gate-input triggering.

volts, rms. The squarewave output signal symmetry varies with the input signal amplitude, so R8 must be adjusted to optimize that symmetry. The Schmitt trigger performs satisfactorily as a sinewave-to-squarewave converter at frequencies up to a few hundred kilohertz. The device produces squarewave output signals whose rise times are only a fraction of a microsecond.

Sawtooth generators

The astable multivibrator shown in Fig. 3 is one of a variety of circuits that can generate sawtooth waveforms. For example, it can generate negative-going sawtooth waves at the bases of both transistors Q1 and Q2. As a result, the astable multivibrator can be considered as another *free-running sawtooth generator*.

Similarly, the monostable multivibrators shown in Figs. 5 to 8 each generate a negative-going sawtooth on the base of Q2 during their active phases. They can be considered as *triggered sawtooth generators*.

Practical versions of Figs. 5 to 8 generate slightly nonlinear sawtooth waveforms because each of their timing capacitors charge exponentially (rather than linearly) through their timing resistors. This aberration can be easily overcome by replacing each timing resistor with a constant-current generator capable of generating linear waveforms.

A timing circuit based on the 555-type integrated circuit

timer offers the best way to generate positive-going triggered sawtooth waveforms. However, if you want to generate free-running, positive-going sawtooth waveforms, this can be done with a unijunction transistor or UJT, connected in the circuit shown in Fig. 12.

The UJT is a three-terminal

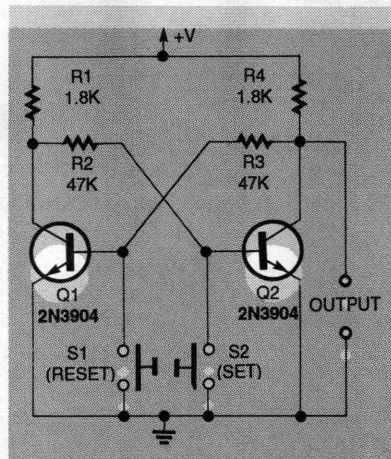


FIG. 9—A SWITCH-TRIGGERED flip-flop (R-S) bistable multivibrator.

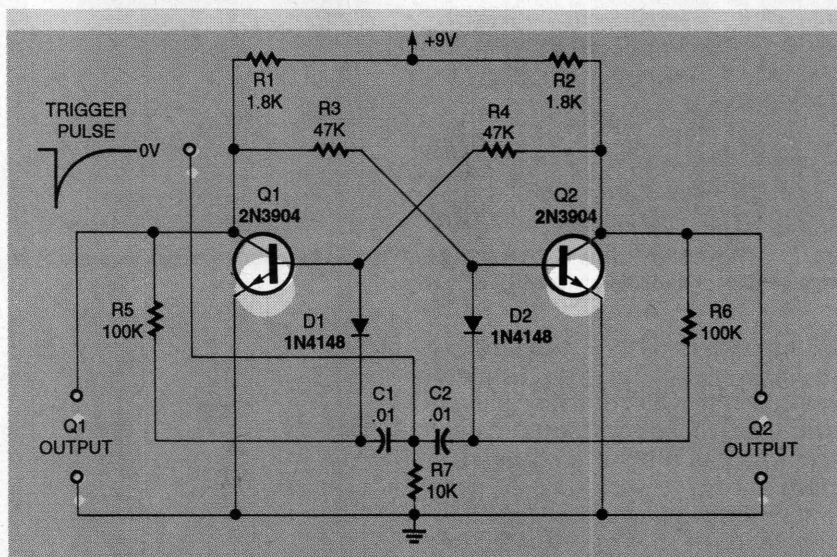


FIG. 10—A DIVIDE-BY-TWO BISTABLE circuit.

transistor whose terminals are identified as *emitter* (E), *base 1* (B1), and *base 2* (B2). A UJT is connected as shown in Fig. 12 as Q1 with its B2 positive with respect to B1, and with the input applied to its emitter terminal.

The emitter of the UJT Q1 presents a very high impedance until the input (emitter) voltage reaches a specific *firing voltage*. At that time, UJT Q1 switches

abruptly to the ON state. When it is on, the emitter presents a low input impedance, and it draws a significant amount of current from the input circuitry. However, if this input current falls below a certain threshold value, UJT Q1 automatically switches back to its high input impedance state.

In Fig. 12, capacitor C1 charges exponentially towards the positive supply voltage through trimmer potentiometer R4 and R1 until the voltage on C1 reaches the firing value of the UJT Q1. At that time, the Q1 switches on and rapidly discharges C1. As soon as C1 is discharged, Q1 turns off again, so C1 starts to recharge again through R4 and R1.

This circuit generates a stable but nonlinear sawtooth waveform that can be varied from 25 Hz to 3 kHz by R4, with the value of capacitor C1 shown. Transistor Q2 and Q3 are connected as a Darlington emitter-follower

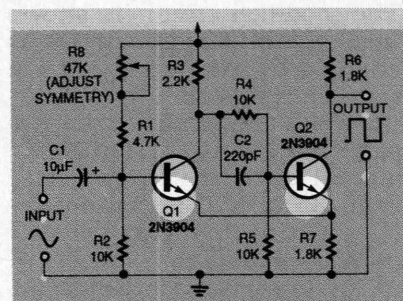


FIG. 11—SCHMITT TRIGGER sinewave-to-squarewave converter.

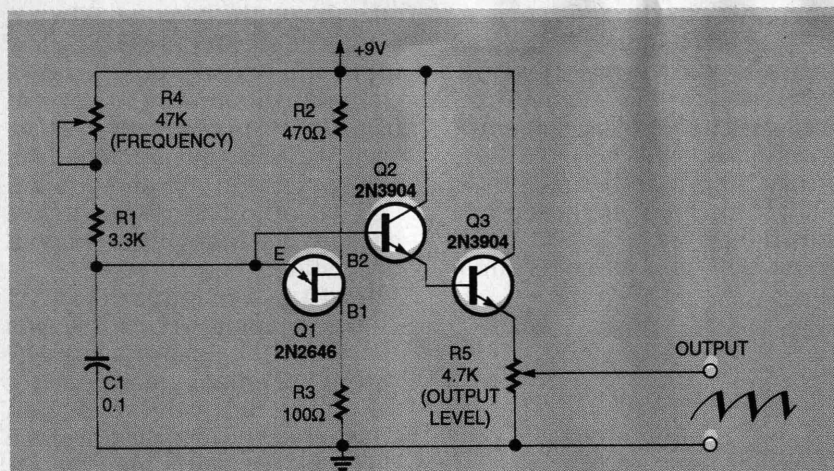


FIG. 12—A NONLINEAR SAWTOOTH GENERATOR that works over a range of 25 Hz to 3 kHz.

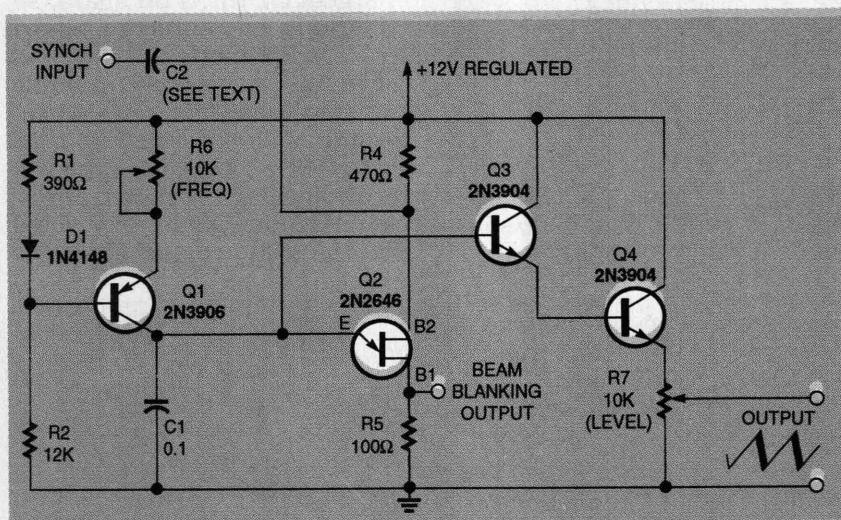


FIG. 13—THIS LINEAR SAWTOOTH GENERATOR can function as a oscilloscope timebase generator and can blank the CRT beam.

buffer stage. This arrangement makes a low-impedance sawtooth waveform available at an output terminal taken from the wiper of output level potentiometer R5.

The linear sawtooth generating circuit in Fig. 12 can be modified to become an oscilloscope timebase generator. The modified circuit is shown in Fig. 13. Capacitor C1 is charged by a constant-current source. In this circuit, Q1 functions as a temperature-compensated, constant-current generator. Its current can be varied from 35 to 390 microamperes by adjusting frequency trimmer potentiometer R6.

The linear sawtooth is available as a variable output whose amplitude can be varied by setting level potentiometer R7. The

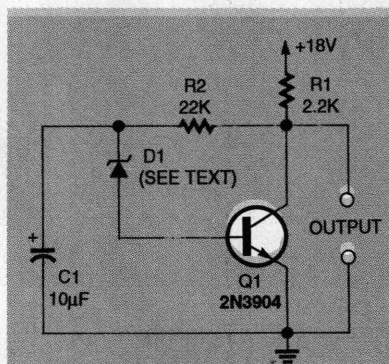


FIG. 14—A WHITE-NOISE GENERATOR has many applications.

output between R7 and ground can be fed via a coaxial cable to the external timebase jack of an oscilloscope.

Positive "flyback" pulses taken between resistor R5 and B1 of UJT Q2 at the beam-blanking output can be used to blank the oscilloscope beam if taken

through a high-voltage blocking capacitor.

The operating frequency of the Fig. 13 circuit can be varied from 60 to 700 Hz with R6 if all of the component values are as shown. Other frequency ranges can be obtained by substituting other values for capacitor C1. The timebase generator can be synchronised to an external signal by feeding the external signal to UJT Q2 through the sync input capacitor C2.

This external signal, which must have a peak amplitude between 200 millivolts and 1.0 volt, effectively modulates the supply voltage (and thus the trigger point) of UJT Q2. It causes UJT Q2 to fire in synchronism with the external trigger signal.

Capacitor C2 must have a lower impedance than resistor R4 at the sync signal frequency. Also, capacitor C2 must have a working voltage that is greater than the external voltage from which the external signal is applied. If the sync signal has a rectangular form with short rise and fall times, the value of C2 need only be a few hundred picofarads.

White-noise generator

"White noise" is another useful waveform. It is a signal that contains a full spectrum of randomly generated frequencies, each having equal mean power when averaged over a unit of time. White noise is useful for testing audio and radio frequency amplifiers, and it is widely used to mask background noise to serve as a sleeping aid.

Fig. 14 is the schematic for a simple, practical white-noise generator. It operates on the principle that all reverse-biased Zener diodes inherently generate white noise. In Fig. 14, R2 and D1 are connected in a negative-feedback loop between the collector and base of common-emitter amplifier Q1. Negative feedback stabilizes the DC working levels of the generator. Capacitor C1 serves to decouple alternating current from the circuit.

The Zener diode acts as a

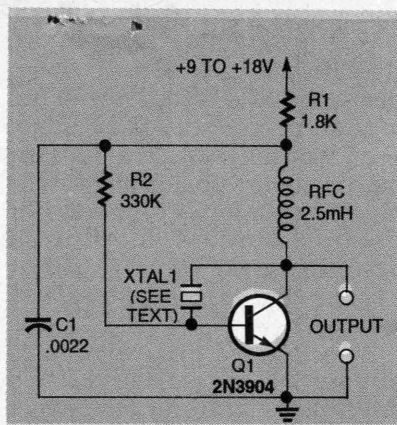


FIG. 15—A PIERCE OSCILLATOR with a parallel-mode crystal.

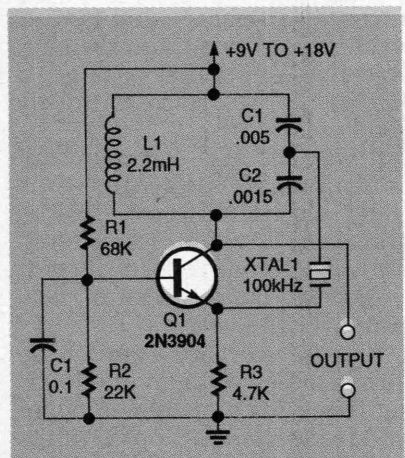


FIG. 16—A 100-kHz COLPITTS oscillator with a series-mode crystal.

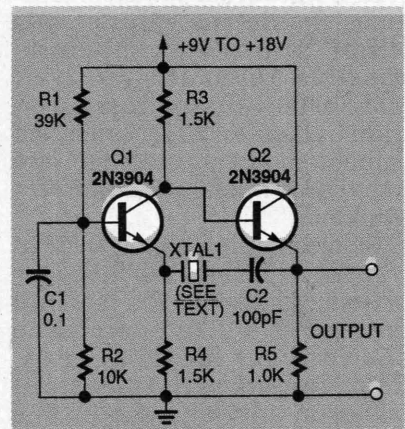


FIG. 17—THIS 50-kHz to 10-MHz oscillator will work with most series-mode crystals.

white-noise source that is in series with the base of transistor Q1. The Zener noise is amplified by the transistor to a useful level of about 1 volt peak-to-peak. Any Zener diode rated for 5.6 to 12 volts should work well in this circuit. Try different Zener diodes and compare the white-noise output.

Crystal oscillators

Crystal oscillator circuits generate accurate, stable frequencies because they include precisely cut piezoelectric quartz crystals which function as high precision electromechanical resonators or tuned circuits. The crystals in these circuits typically have Q s of about 100,000, and they can provide as much as 1000 times greater frequency stability than can conventional inductive-capacitive (LC) tank-circuit oscillators.

A piezoelectric crystal's operating frequency of a few kHz to 100 MHz is determined by its mechanical dimensions. The crystal can be cut to provide either series or parallel resonant operation. Series-mode crystals present a low impedance at resonance, while parallel-mode crystals present a high impedance at resonance.

Figure 15 is a practical schematic for a crystal oscillator that is designed for a parallel-mode crystal. The circuit is actually a Pierce oscillator, and it will oscillate with most 100-kHz to 5-MHz parallel-mode crystals without any circuit modification.

Figure 16 shows an alternative 100-kHz oscillator that was designed for a series-mode crystal. It is known as a Colpitts oscillator.

Its tank circuit, consisting of L1, C1, and C2, is designed to resonate at the same frequency as the crystal. However, the tank circuit component values must be changed if any other crystal frequencies are desired.

Figure 17 is the schematic for a useful two-transistor oscillator that will work with most 50 kHz to 10 MHz series-resonant crystals. In this circuit, Q1 is connected as a common base amplifier, and Q2 is an emitter follower. The output signal (from Q2's emitter) is fed back to the input (Q1's emitter) through C2 and the series-resonant crystal. This is a versatile oscillator circuit that will work even with a low-cost, marginal crystal. Because of that, the circuit can form the heart of a simple crystal tester.

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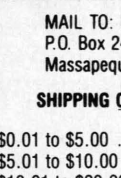
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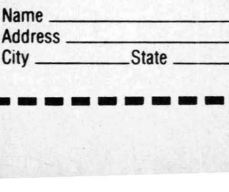
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We're finally free to play our game without performing the document check!

ROBERT GROSSBLATT

If you've been following this series on software debugging (it's been discussed in the July, September, and November issues of **Electronics Now**), you now know the basic steps to getting rid of a document check. The job involves locating the relevant code and getting rid of it. Although that sounds simple enough, finding the code can be a difficult procedure—especially when there aren't any roadsigns to follow.

The pages and pages of code generated by raw debug dumps is, without question, some of the most stultifying and mind-numbing text you'll ever read. But after a bit of experience, the relevant portions will jump off the screen at you—even though the people who write document-check or copy-protection routines often go out of their way to make it as difficult to find and follow as possible. But more about that later. Let's first get back to finishing off our example.

Recall the steps taken so far. First, a list was made of exactly what happens when the game is played, and the order in which those things happen. We also noted what happens when a document check is answered correctly and what happens when an incorrect answer is given. Then the CALLs that initiated those routines were found and carefully examined for any clues as to the operation of the document check. In our example I've created a JUMP table that ran one part of the game's introduction after another. To refresh your memory, the disassembly listing for the code we're working on is shown in Listing 1. While things probably won't be quite as neat as Listing 1 in an actual program, the basic idea will be the same.

Simply eliminating the document check is usually a bad idea. Putting a bunch of NOPs (no operation) in

place of the CALL to the document check will certainly bypass the routine but, as we saw last month, it can also cause the game to crash. That's because when the document check is bypassed, anything that it was supposed to do—such as passing data to the main program—is bypassed as well. If the data is found by the main program, the game gets started. If not, you're out of luck.

Tracing into the document check last month revealed that entering the correct code word caused the routine to set a flag, which allowed the game to continue normally. In the disassembled code shown in Listing 2, the key instruction is the last one before the return. If a 1111h is put in location 1C32h, the game can be started by executing the call that starts the game. What we have to do is figure out how to make that happen automatically.

As with most things, there are lots of ways to do the job. In our example, the most straightforward

way to do it is to take the instruction at 554F:62BA and substitute it for the CALL found at 2BC0:190C—the one that calls the document check in the first place. By doing that, the program will set the flag correctly and then go on to execute the game. The only problem with this is that we don't have enough room in the jump table, because the MOV instruction is six bytes long and the CALL is only five bytes.

That's not much of a problem because we actually have a huge area of unnecessary code that we can overwrite—the document-check code itself. Once we finish patching the code so that it will bypass the document check completely, none of the code there will ever be executed. That is, after all, the reason behind this whole exercise. The document check starts at location 554F:5D96 (see the third line in Listing 1). That's the address where the instruction to set the flag and return to the jump table should be patched. The actual code we need

LISTING 1

2BC0:1902	9AB3194F55	CALL 554F:19B3	;Show the company logo
2BC0:1907	9AC6304F55	CALL 554F:30C6	;Play some music
2BC0:190C	9A965D4F55	CALL 554F:5D96	;Do a document check
2BC0:1911	9A3ADB4F55	CALL 554F:DB3A	;Go to the game

LISTING 2

554F:629C	837EDE00	CMP WORD PTR [BP-22],+00
554F:62A0	7402	JZ 62A4
554F:62A2	EB0C	JMP 62B0
554F:62A4	FF46D6	INC WORD PTR [BP-2A]
554F:62A7	837ED602	CMP WORD PTR [BP-2A],+02
554F:62AB	7D03	JGE 62B0
554F:62AD	E937FD	JMP 5FE7
554F:62B0	837EDE00	CMP WORD PTR [BP-22],+00
554F:62B4	7504	JNZ 62BA
554F:62B6	0E	PUSH CS
554F:62B7	E8E8FC	CALL 62A2
554F:62BA	C706321C1111	MOV WORD PTR [1C32],1111
554F:62C0	CB	RET

is the last two instructions found at the very end of the document check itself, that is, the last two lines in Listing 2. The program will execute normally, go to the document check, set the flag correctly, and execute the game.

This is easy enough to try while still inside the debugger. All we have to do is edit the instructions at the beginning of the document check and replace whatever we find there with those two instructions. As soon as that's done, the debugger's JUMP command can execute the start of the entire game at the top of the original jump table.

Making this change permanent is a piece of cake. All you need is a hexadecimal editor, a bunch of hex bytes, and a search string. The first is a matter of preference and convenience, the second is the hex equivalent of the last two instructions in the document check, and the third is nothing more than the original hex code found at the beginning of the document check.

Use your hex editor's search command to find the beginning of the document check, patch in the seven replacement bytes, and write the file back to disk with those changes. Once you do that, you should be able to go back to the command prompt, enter the name of the EXE file, and run the game. Everything will be as it was except the document check will be gone.

The example presented here is much easier than what you're likely to encounter in a real program. Since the job of any copy-protection routine is, as the name would imply, to protect against copying, the code that does the work is often hidden and hard to find. You're undoubtedly going to run across techniques such as self-modifying code and hidden checks that can make a complete crack of the program more than you bargained for. In our example, the crack could have been accomplished by something as simple as changing the address in the CALL to the document check. Instead of adding a patch to the beginning of the document check, we could have simply CALLED the end of the check where the flag was set by the original code. In that case, the jump table would have been

LISTING 3

```
2BC0:1902 9AB3194F55 CALL 554F:19B3 ;Show the company logo
2BC0:1907 9AC6304F55 CALL 554F:30C6 ;Play some music
2BC0:190C 9ABA624F55 CALL 554F:62BA ;Do a document check (HA!)
2BC0:1911 9A3ADB4F55 CALL 554F:DB3A ;Go to the game
```

LISTING 4

```
1BCE:01D2 8A0E9000 MOV CL,[0090]
1BCE:01D6 E8BEFF CALL 0197
1BCE:01D9 380E9000 CMP [0090],CL
1BCE:01DD 74F3 JZ 01D2
1BCE:01DF A08000 MOV AL,[0080]
1BCE:01E2 FEC0 INC AL
1BCE:01E4 3C0A CMP AL,0A
1BCE:01E6 7405 JZ 01ED
1BCE:01E8 A28000 MOV [0080],AL
1BCE:01EB EBE5 JMP 01D2
1BCE:01ED C606800000 MOV BYTE PTR [0080],00
1BCE:01F2 A08100 MOV AL,[0081]
1BCE:01F5 FEC0 INC AL
1BCE:01F7 3C0A CMP AL,0A
1BCE:01F9 7405 JZ 0200
1BCE:01FB A28100 MOV [0081],AL
1BCE:01FE EBD2 JMP 01D2
1BCE:0200 C606810000 MOV BYTE PTR [0081],00
1BCE:0205 A08200 MOV AL,[0082]
```

LISTING 5

```
2A23:011F 52 PUSH DX
2A23:0120 BACE1B MOV DX,1BCE
2A23:0122 C706D201B39A MOV WORD PTR [01D2],9AB3
2A23:0128 C706D4014F19 MOV WORD PTR [01D4],194F
2A23:012E C706D6019A55 MOV WORD PTR [01D6],559A
2A23:0134 C706D80130C6 MOV WORD PTR [01D8],C630
2A23:013A C706DA01554F MOV WORD PTR [01DA],4F55
2A23:0140 C706DC01BA9A MOV WORD PTR [01DC],9ABA
2A23:0146 C706DE014F62 MOV WORD PTR [01DE],624F
2A23:014C C706E0019A55 MOV WORD PTR [01E0],559A
2A23:0152 C706E201DB3A MOV WORD PTR [01E2],3ADB
2A23:0158 C706E401554F MOV WORD PTR [01E4],4F55
2A23:015E C706E6010E5A MOV WORD PTR [01E6],5AB8
2A23:0164 C706E8010700 MOV WORD PTR [01E8],0007
2A23:016A B000 MOV AL,00
```

changed to look like Listing 3, and the result would have been the same as it was when we patched the beginning of the document-check code.

The bottom line in defeating copy-protection schemes is that you have to work under the assumption that nothing should be taken at face value. If you've plodded through the code tracing every CALL and haven't found the one that brings up the document-check screen, there's a good chance that the actual CALL is hidden in some self-modifying code. That can be a real pain

in the neck to find because the code being modified can appear to be anything at all. It can look perfectly legitimate and even the most astute crackist can be fooled.

As an example, what's wrong with the code in Listing 4? The answer is that there's nothing wrong with it; it's part of a routine that I wrote some years ago for an EPROM in a microprocessor-controlled lighting system. If you came across this in a game you wouldn't give it a second glance.

Taking this a step further, suppose you ran across code that

LISTING 6

```

1BCE:01D2 9AB3194F55 CALL 554F:19B3 ;Show the company logo
1BCE:01D7 9AC6304F55 CALL 554F:30C6 ;Play some music
1BCE:01DC 9ABA624F55 CALL 554F:62BA ;Do a document check
1BCE:01E1 9A3ADB4F55 CALL 554F:DB3A ;Go to the game
1BCE:01E6 5A POP DX ;More code
1BCE:01E7 OE PUSH CS ;And more code
1BCE:01E8 A28000 MOV [0080],AL
1BCE:01EB EBE5 JMP 01D2
1BCE:01ED C606800000 MOV BYTE PTR [0080],00
1BCE:01F2 A08100 MOV AL,[0081]
1BCE:01F5 FEC0 INC AL
1BCE:01F7 3C0A CMP AL,0A
1BCE:01F9 7405 JZ 0200
1BCE:01FB A28100 MOV [0081],AL
1BCE:01FE EBD2 JMP 01D2
1BCE:0200 C606810000 MOV BYTE PTR [0081],00
1BCE:0205 A08200 MOV AL,[0082]

```

looked like Listing 5. That is perfectly normal code as well. If you happened to come across it, especially if it was surrounded by a bunch of other similar MOV instructions, you might assume that something was being set up such as music, sound effects, graphics, or whatever. Actually though, it is some sneaky code whose execution results in overwriting the first piece of innocent code so that it will now look like Listing 6.

Listing 6 is, of course, the jump table we've seen before. It doesn't exist until after the self-modifying code has been executed. A trick that's even more difficult to decipher is using obscure routines to calculate the bytes. Anything can be done by the programmer, whose rule is to make the copy-protection technique as hidden and arcane as possible.

In general, if you can't find the jump table or the CALLs to the major routines in a piece of software, you're probably up against self-modifying code. The best approach is still to find the CALLs that run the routines that you can recognize in the game. Those include such things as playing music and putting messages on the screen.

Once you see a connection between sections of code and different events in the game, you can work backwards to locate other signposts. It's not easy—it's designed to be as difficult as possible. The only piece of encouragement I can offer to you is that the more you do this, the better you'll get.

You won't find any ready-to-use software that can do this job for you, and you won't find a shelf full of books to help you either. You've got to do it on your own and, as an extra side benefit, the better you get at deciphering copy-protection schemes, the better your own programming skills will become.

When we get together next time, we'll get back to circuits. Ω

AUDIO UPDATE

continued from page 80

But aside from those cavils, I found Davis's conclusions to be right on target. He found the differences between two-wire cables to be indistinguishable. Comparing two-wire cables with flat ribbon cables revealed a slight difference in the high treble averaging about 0.5 dB at 15 kHz. These differences are at the threshold of audibility and are a far cry from the "quantum leaps" of improvement extolled by the manufacturers and the audiophiles they've entranced.

Davis confirms my view that unless strange amplifier or speaker impedances are involved, 12-gauge "zip cord" will work fine. He concludes that the use of exotic materials for the conductors, or strange cable insulation, or oxygen-free copper multi-strand "hyperlitz" windings will have minimal positive effect on the signal, but maximal negative effect on your wallet. Ω

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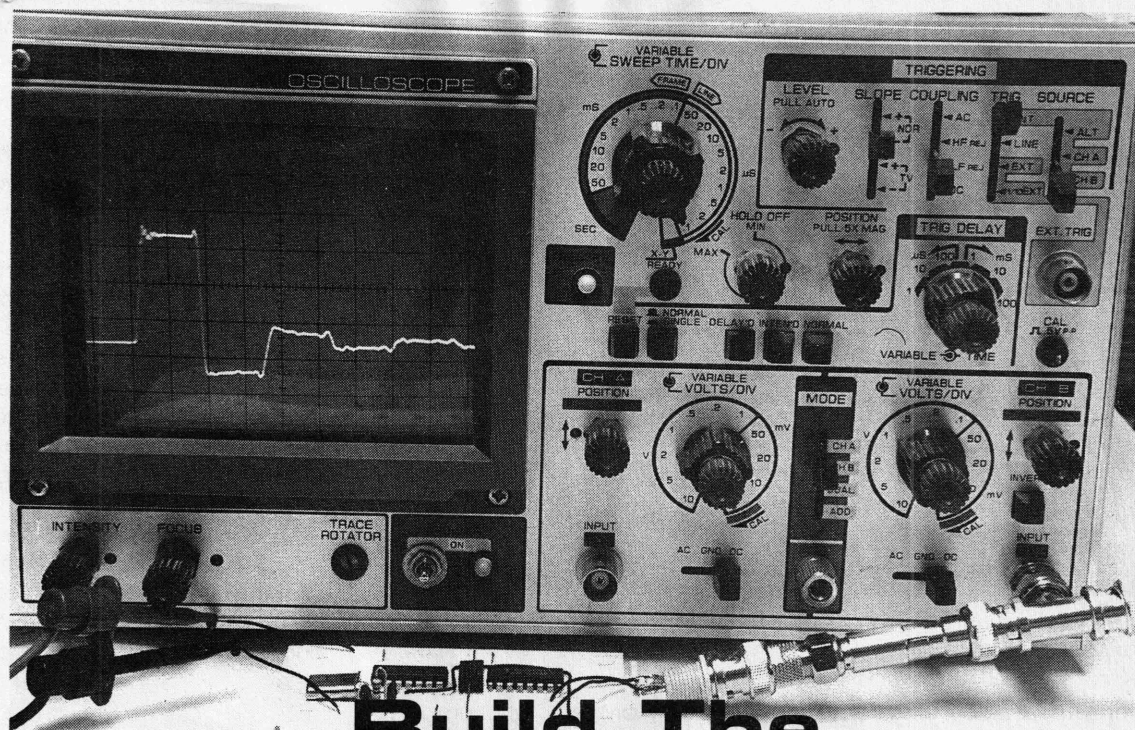


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Need to test a damaged or suspect coaxial cable? This handy circuit will do the trick without sending you to the poorhouse for buying expensive equipment. If you supply the oscilloscope and about five bucks for parts, we'll supply the know-how.

JAMES E. CICON

You have a spool of coaxial or twin-lead cable and want to know how long it is. Unwinding the cable would be impractical.

After a rainstorm, you sometimes lose the picture on your television set. You figure that the satellite-TV antenna cable probably has a crack in the insulation somewhere along its length. The problem is that the cable is over 250 feet long. It snakes through your walls, across your roof, and then drops down underground for the run to the dish. It would be nice to know how far down the line the crack is.

Your local-area network at work goes down whenever a certain machine is brought on-line. The installation techs say that the LAN administrator has the software configuration fouled up. The LAN administrator says the installation techs had been work-

ing too much overtime when they hooked up that particular machine. Can you pinpoint who is right?

Televisions, computer networks, CBs, ham radios, cellular telephones—all of those devices send or receive information using electrical waves that might travel through a cable for at least some distance. If the cable is damaged, defective, or the wrong type for the job, the original signal traveling down the cable will become garbled. That will cause poor picture reception on a TV, data collisions on a LAN, and other problems. Wouldn't it be great if you could easily and quickly diagnose the cause of those types of cabling headaches?

There is an instrument that can do just that. It's called a "Time Domain Reflectometer," or TDR. A TDR can tell you many things about the health of a cabling system. It can quickly and easily tell you the length of cable. It

can also tell you if there is an impedance mismatch (such as two different types of cable spliced together) on a cable. It can help you locate a fault in a cable such as a break or a short circuit. A TDR has only one real drawback—its price tag. You can easily pay thousands of dollars for one.

But there is a less expensive alternative. In this article we're going to show you how to build and use the Cable Reflection Tester, or C.R.T. for short—a device that will let you see the effects of the various cable faults mentioned before. Not only does the C.R.T. demonstrate some very interesting properties of electrical waves from a purely academic standpoint, but the project has a very practical side. We'll take you through a few experiments to demonstrate how to use the C.R.T. to measure the length of a cable, and to determine the health of a cable.

Using the C.R.T. requires an oscilloscope, and a handful of inexpensive parts for the C.R.T. itself. If you don't already have most of the parts on hand, you can probably buy them for less than \$5. The prototype was originally built and tested on a solderless breadboard. If you really like the C.R.T., there is no reason not to mount it permanently inside the oscilloscope. You could then tie it into the oscilloscope's power supply and be able to use it wherever and whenever you wish.

A fancy scope is not needed for the C.R.T. The scope used here is a beat-up 35-MHz model that is over ten years old. It has been through a flood and dropped several times. Despite the abuse, it works surprisingly well with the C.R.T.

Background Theory. The time-honored way to understand how electricity works is to think of electricity as water and wires as pipes. That mental image has helped many students to understand electrical phenomena, and is also one that is very useful in explaining how the C.R.T. works. The analogy is not perfect and breaks down if pushed too far, but for our needs, it still has a lot of value.

Imagine spilling a glass of water into a tray. The water will start to spread out over the bottom of the tray until it comes to a side. While it is spreading, the puddle will have some depth. That depth depends on how fast the water is being poured into the tray and how fast the water can spread out to fill the tray. Water clings better to some surfaces than to others. If you are pouring the water into a tray made of a material to which water clings, the water will spread more slowly, and the resulting depth of the puddle while it is spreading out will be deeper.

That "clinginess" is a feature of the material of which the tray is made. It "impedes," or restricts, the flow of the water. You could call the clinginess of a tray its *characteristic impedance*. How fast the water spreads across the tray (its *velocity factor*) largely depends on the tray's characteristic impedance. To compare the velocity factor of one tray to another, you could pick one tray as a standard. If water spreads half as fast in one tray than in the tray that is the standard, you could say that the velocity factor

of the tray under test is 0.5 or 50%.

What happens when the water reaches the side of the tray? It will slosh back, of course. Ripples will travel back from the side to the center where the water is being poured into the tray. The level of the water in the tray will now start to rise. Remember that before the water reached the side of the tray, the level of the water was determined only by how fast the water could spread out. Now the level of the water is determined by how much water you pour in. Another way to think of it is that, at first, the water level is determined by the charac-

teristic impedance of the tray. Time how long it took for the water to start rising from the moment you started pouring the water in. That is the time that it took the water to reach the side of the tray plus the time it took for the returning wave to bounce back to the center.

Divide the time in half, and then multiply that result by the tray's velocity factor. That gives the distance from the center of the tray to its nearest side. If you are wondering why you divided the time in half, it is because the water traveled out to the side of the tray and then back again, actually traveling twice the distance you want to measure.

We can take that procedure and express it as an equation:

$$L = t/2 \times V_f \times V_r$$

where L is the distance to the side of the tray, t is the time for the wave to travel to the side of the tray and back again, V_f is the tray's velocity factor, and V_r is the velocity of a wave traveling in the reference tray.

How the C.R.T. Works. The C.R.T. works in a very similar way to the example of pouring water into a tray. Substitute the pouring water with a signal generator and the tray with a cable. Instead of watching the water level with your eyes, use an oscilloscope to help you monitor the electrical voltage level on the cable. Figure 1 shows the basic block diagram of the C.R.T.

When a squarewave from the signal generator goes high, you will see that on the oscilloscope. Notice that the voltage level doesn't immediately rise to the level that the signal generator is producing. It initially steps to some intermediate value. After a short delay, the level changes to either a higher level or a lower level, depending on whether the far end of the cable is open- or short-circuited. That will show up as a stair-step waveform on the oscilloscope.

Let's think about why that happens. The first step in the waveform is equivalent to just starting to pour the water into the tray in our example. The voltage level is determined largely by the characteristic impedance of the cable. If the end of the cable is open, the voltage wave has nowhere to go when it reaches the end of the line. That is the electrical equivalent of hit-

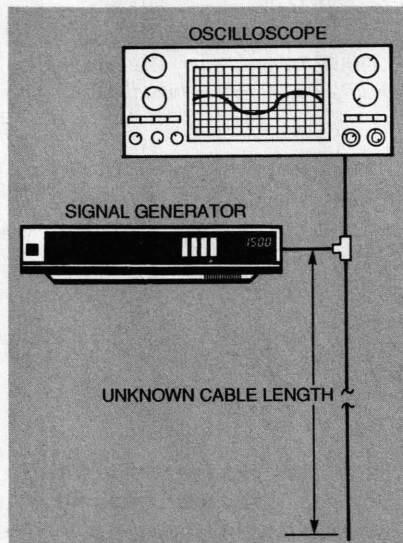


Fig. 1. This setup will let you easily measure the length of a cable without using a tape measure. Connect a signal generator to one end of the cable and measure the amount of time it takes to see the steps in the waveform on an oscilloscope.

teristic impedance of the tray. After all the waves and ripples have settled down, the level is determined by what the original potential of the water was to fill the tray.

Let's say that you want to know the distance from the center of the tray to the side of the tray. However, you are limited to two measurements: the height of the water in the center of the tray, and time. You also know the tray's velocity factor.

To find out how far the side of the tray is from the center where you are pouring water in, you could start pouring water into the center. Keep an eye on the height of the water in the center of the tray. The water height will stay constant for a while and then

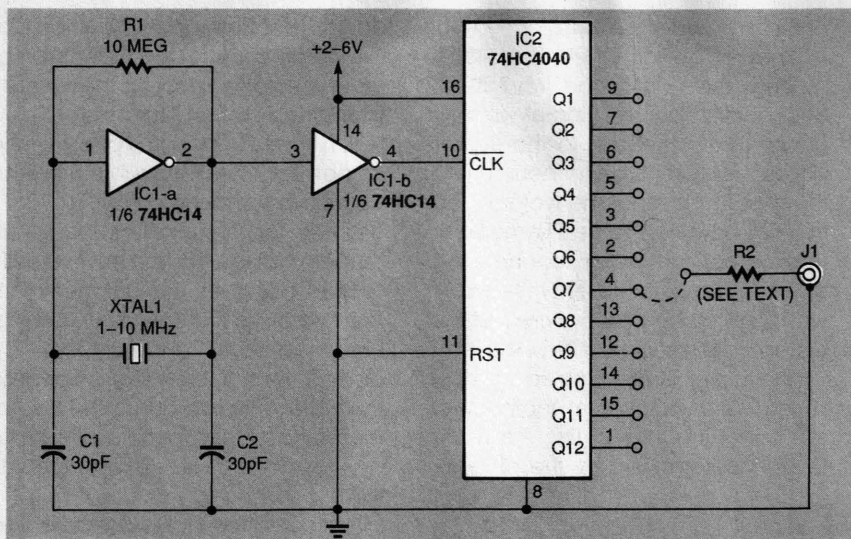


Fig. 2. Here's the schematic diagram for the Cable Reflection Tester. The components shouldn't cost more than \$5, and it can be built up on a solderless breadboard in only a few minutes.

TABLE 1

Cable Type	Velocity Factor	Characteristic Impedance
RG-58/U	66%	52 ohms
RG-59/U	66%	75 ohms
RG-6/U	75%	75 ohms
RG-8/U	66%	52 ohms
RG-8/M	75%	52 ohms
RG-174	66%	50 ohms
RG-62A/U	84%	93 ohms

ting the wall of the tray and the voltage wave is reflected back, which is seen as a second step up as it reaches the scope. Eventually, after all the ripples settle down, the cable "fills" up, and the final voltage level is equal to the level being produced by the signal generator.

If, on the other hand, the end of the cable is shorted, something different happens. The water analogy does not work as well as when the cable is open. As the signal generator drives the line high, the initial step up will appear on the scope as the voltage wave starts to travel down the cable. When the wave gets to the end of the cable, it sees a short circuit. Electrically speaking, the short circuit acts like a reverse wall. The wave striking it is bounced back upside down, canceling out some or all of the forward-traveling wave. When that negative-traveling wave finally travels back to the scope, it will be seen as a step down.

Those stair-step waveforms can be used in many different ways. You can measure the width of the first stair step to determine the length of the cable.

Remember that the width of the first stair step is the time it takes the initial wave to travel to the end of the line and back, just like measuring the length of the tray in the water analogy. The other thing you can determine is if the end of the line is open or shorted. We'll do those experiments later. First, we need to build the C.R.T. itself. Generating and viewing the waveforms on an actual cable is a lot better than just talking about them.

Designing and Building the C.R.T.

The C.R.T.'s design is so simple, building it is almost trivial. All we need is a squarewave signal generator that has a very short rise time. A crisp square-wave will result in nice, sharp stair-step display, making it easier to see the effects that we're looking for. The schematic in Fig. 2 shows how simple the design actually is.

Integrated circuit IC1-a is an inverter. It is the active component in an oscillator that generates the C.R.T.'s squarewave. The rest of the oscillator consists of R1, XTAL1, C1, and C2. Resistor R1 provides a DC feedback path for IC1-a. The value of R1 is not critical, but should not be too large. The suggested value of 10 megohms will work quite well. Capacitors C1 and C2 stabilize the oscillator, and their values are also not critical, but should be between 10 and 60 pF. The smaller their value is, the more quickly the oscillator will start, but the less stable it will be. XTAL1 sets the frequency of the oscillator. Its value is also not critical, but should be between 1 and 10 MHz for good results.

The output of IC1-a is buffered by IC1-b and fed to IC2, whose purpose is to divide the signal frequency down so that cables of various lengths can be measured. Without IC2, the reflected waves in very long cables may not return in one complete oscillator cycle. By slowing the clock frequency down with IC2, it is easier to measure longer cables. If you are only working with short cable runs, it might be possible to omit IC2.

The oscillator is matched to the cable being tested with R2. Its value should be equal to the characteristic

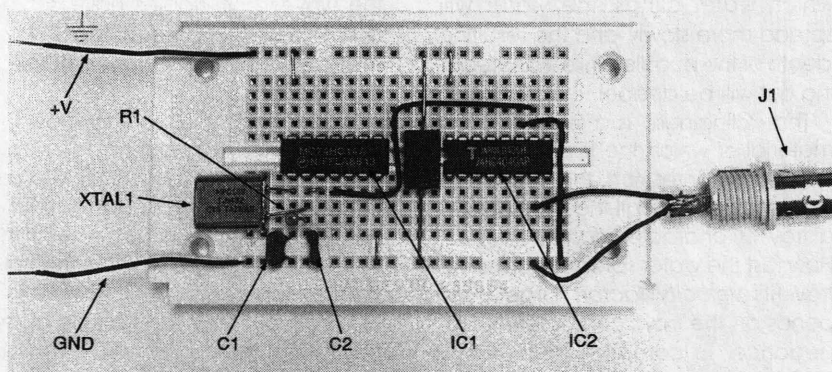


Fig. 3. If you want to build the C.R.T. on a solderless breadboard first, you can follow this photo as one way to locate the components. Resistor R2 is not being used in this version of the C.R.T., and an additional .01-0.1-μF capacitor has been added between IC1 and IC2 to smooth out the power supply from ripples caused by the integrated circuits.

impedance of the cable being tested. That match helps prevent multiple reflections, and helps make the stair-step response easier to see. Resistor R2 is not absolutely needed, especially if you are concerned with only the first step of the stair-step waveform.

All of the components for the C.R.T. can be mounted on a small breadboard. The photograph in Fig. 3 shows both the completed prototype and placement of the parts. Assembly can be finished in 5 minutes or less. The power supply for the circuit can be anywhere between 2 and 6 volts, so the C.R.T. can be powered with your choice of either batteries or a bench-top power supply.

Experiments. We've already seen (from the water-tray example) a method that allows us to measure the distance the water travels to the side of the tray. That method can be applied to electrical waves traveling in cables as well. An oscilloscope can be used to measure both the time that a wave takes to travel the length of a cable, and what the voltage level of that wave is. Some common types of coaxial cable are listed in Table 1 along with their velocity factors and characteristic impedances.

An electrical wave's velocity factor (how fast the water spreads across the tray in our analogy from before) is related to the speed of light in a vacuum. For example, according to the

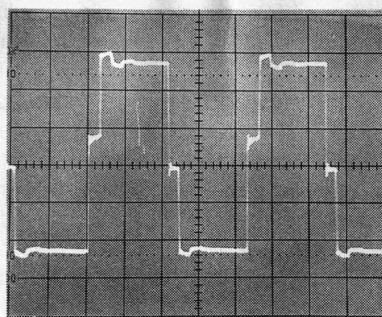


Fig. 4. Here's the C.R.T. hooked up to 50 feet of RG-59/U coaxial cable. The plateau halfway up the pulse is the effect of the signal bouncing back from the end of the cable. The length of time of that plateau measures how long it took for the signal to travel to the end of the cable and back.

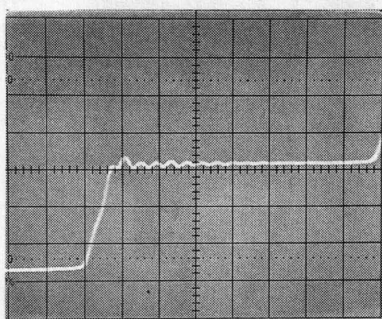


Fig. 5. In this photo, the scope is magnifying the plateau mentioned in Fig. 4. Since we know how fast electricity travels in this particular cable, half the distance shown here is how long it takes for the signal to go from one end of the cable to the other. It's now a simple matter to apply $\text{Distance} = \text{Time} \times \text{Speed}$ to figure out the length of the cable.

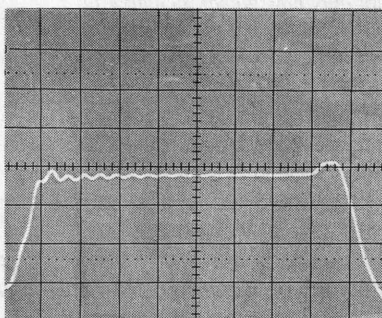


Fig. 6. Now we've shorted the far end of the cable. Instead of going higher, the reflected signal makes the display drop down. Not only does that tell us that the cable is shorted, it also can tell us the distance to the short.

information in Table 1, a wave in RG-58/U cable travels at 66% of the speed of light. We can use the

characteristic impedance of the cable to tell us what value we should use for R2.

Now for some experiments. Connect the C.R.T. to an oscilloscope as shown in Fig. 1 using a 50-foot length of RG-59/U coaxial cable. Leave the long end of the cable unconnected. The waveform on the scope should be similar to that shown in Fig. 4.

You can easily see the very obvious stair-step effect in both the leading and trailing edges of the squarewave. Figure 4 is interesting because it shows several complete cycles of the squarewave and the overall effect of the reflected wave upon it.

In order to get a useful measurement however, you must zoom in on just the first step on either the leading or trailing edge. Figure 5 shows the waveform of Fig. 4 with the scope's magnifier turned on. That will, unfortunately, decrease the accuracy of the measurement. Count the divisions on the scope's screen. The example in Fig. 5 has a stair-step about 8 divisions wide. Make sure that you don't just measure the width of the step itself (which is only 7 divisions wide). Include time it took to rise up to that step (another division). The scope in Fig. 5 is set to .02 $\mu\text{s}/\text{division}$. The velocity factor from Table 1 for the cable (type RG-59/U) is 66%.

Now we can pull out our equation and plug in the numbers. A stair-step of 8 divisions at .02 $\mu\text{s}/\text{division}$ is .16 μs long. Dividing that by 2 gives us .08 μs . We then multiply by both the cable's velocity factor (66%, or .66) and the speed of light, which is 186,284 miles/second. Since our time measurement is in μs , we need to change the speed-of-light constant to μs , also. That conversion alone would be fine if we needed the length in miles, but the cable being tested is only 50 feet long, so multiply the speed-of-light constant by 5280. The final speed-of-light constant we will use in the length formula is now how many feet light travels in 1 μs , which is about 983 feet.

Let's substitute into the formula:

$$\begin{aligned} L &= t/2 \times V_f \times V_l \\ &= .08 \times .66 \times 983 \\ &= 51.9 \end{aligned}$$

The distance of the cable works out to be 51.9 feet. Considering the second-rate accuracy of the scope being

PARTS LIST FOR THE CABLE REFLECTION TESTER

SEMICONDUCTORS

IC1—74HC14 hex HCMOS inverting Schmitt trigger, integrated circuit
IC2—74HC4040 HCMOS 12-stage binary counter, integrated circuit (optional, see text)

RESISTORS

(All resistors are 1/4-watt, 5% units.)
R1—10-megohm
R2—optional, see text

ADDITIONAL PARTS AND MATERIALS

C1, C2—30-pF ceramic-disc capacitor
J1—BNC female connector (Digi-Key ARFX1063-ND or similar)
XTAL1—1-10 MHz crystal, see text
Printed-circuit board or breadboard, wire, battery, etc.

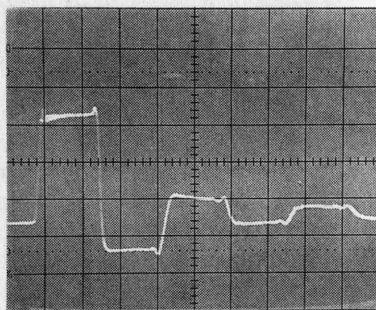


Fig. 7. If you don't use the proper value for R2, the reflected signal will keep echoing back and forth in the cable.

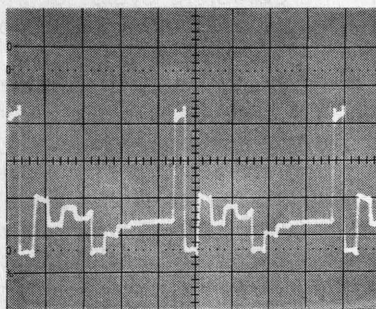


Fig. 8. Multiple echoes and reflections caused by bypassing R2 will mix with the original signal, distorting it badly.

used and the simple carpenter's tape measure used to check the actual length of the cable, the results are surprisingly accurate.

What happens if the end of the cable is shorted instead of left open? The results of that are shown in Fig. 6. Notice that instead of a stair-step waveform, the waveform steps up, then down. The down step is the result of the wave being reflected off of the shorted end of the cable. An additional change to the setup used for Fig. 6 is that R2 has been bypassed. That means that the signal generator is not matched to the cable, so when the reflected wave comes back it is once more reflected out. That back and forth reflection continues on and on like a pendulum with the waveform getting smaller each time. Additional views showing the reflected waves can be seen in Figs. 7 and 8.

Practical Applications. While digging in your yard, you accidentally severed a length of your satellite-TV cable. You had a repairman come out to replace it, but your reception has been terrible since then. The cable in your system is RG-58/U. You suspect that the repairman may have

put in RG-59/U by mistake. How would you use the C.R.T. to diagnose that problem?

To simulate that situation, make up a cable from two 25-foot lengths of RG-58/U cable with a 25-foot length of RG-59/U cable spliced in between them. The overall cable length is 75 feet. Attach the C.R.T. to one end of the cable assembly and watch what

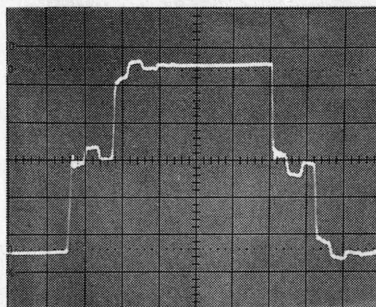


Fig. 9. If there is an impedance mismatch in the cable, the plateau will have an odd "hiccup" as the signal changes speed.

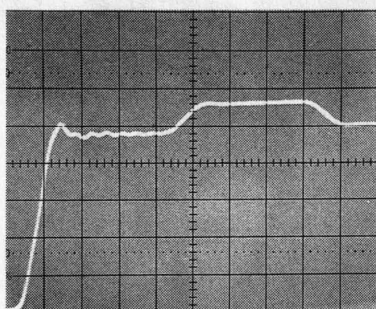


Fig. 10. This magnified view of a mismatched impedance can easily tell you not only how far up the cable the mismatch is, but the length of the offending portion of the cable.

happens on the scope when a squarewave is sent down the cable. The scope in Fig. 9 is set to display one complete cycle of the squarewave that was driving the cable.

The squarewave will initially step up as it travels down the RG-58/U. However, 25-feet later, it hits the RG-59/U. Since RG-59/U has a higher characteristic impedance than RG-58/U, a small positive reflection occurs. That causes the waveform to step up higher. The waveform travels another 25 feet when it hits RG-58/U again. The impedance mismatch (a mirror image of the first mismatch) causes a small negative reflection, which results in the step back down to the val-

ue appropriate for RG-58/U. The waveform continues on to the end of the line, and is then reflected back.

The scope has been set in Fig. 10 to zoom in on the leading edge of the stair-step waveform. If you don't know where the RG-59/U was spliced in, you can easily calculate the location using the length calculations from before. The first part of the waveform is about 4 divisions long. That is the initial length of cable before the splice, which works out to be about 25.9 feet.

The length of the spliced cable causes the short step up, which is also

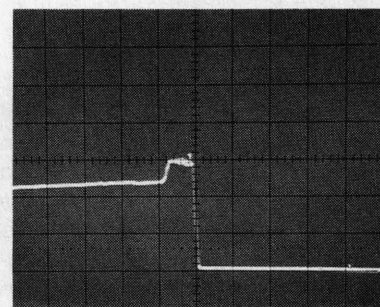


Fig. 11. A step down usually means an impedance mismatch, but the mismatch shown here is a problem of extreme proportions: using the wrong impedance cable! In this test of a computer-network cable, using the wrong type of cable results in garbled data that the interface card cannot read.

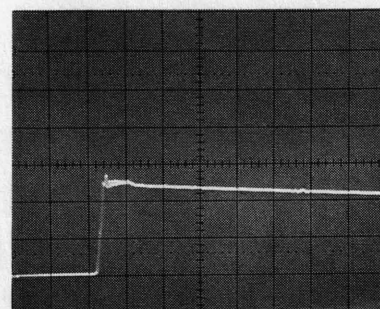


Fig. 12. After the computer-network cable in Fig. 11 is replaced with the proper type, the signal becomes much cleaner. The slow droop in the signal is caused by the capacitor-like quality of the cable: two conductors separated by an insulator.

about 4 divisions long. Since RG-58/U has the same velocity factor as RG-59/U, the same calculations as before can be used, so the length of the spliced cable is also about 25.9 feet.

Don't lose sight, however, of the most important part of that measure-

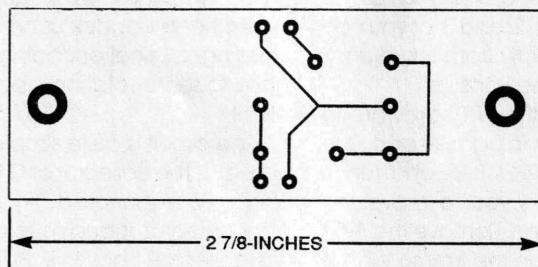
(Continued on page 68)

You can make an encoder disc from poster paper or other stiff, thin material, as shown in Fig. 5. The diameter of the center hole should be slightly smaller than the shaft of the stepper motor. To mount the disc, remove the coupler on the stepper motor and loop a small rubber band around the shaft several times.

Referring to Fig. 6, push the encoder disc onto the shaft. The disc should have a firm press-fit, without bending or distorting. Reinstall the coupler and adjust the disc and the rubber band so that the disc rests directly against

routine. Line 10 senses when pin 11 of J1 goes low (input = 1), indicating that IR energy from the diode is passing through the aperture and pulling the detector's output low. Since the aperture could be detected during any of the four motion steps, line 16 repositions the stepper motor at the start of the sequence (ainit). The program then lets you align the aperture beneath the diode. Last, line 22 de-energizes the motor.

What's Next? There are several directions you could take if you wanted



Since parts alignment is fairly critical, use this PC board when building the optical encoder.

the bottom of the coupler and is held tight by the rubber band. Position the encoder PC board against the short ends of the blocks holding the stepper motor (shown last time), and adjust the positions of the IR emitter and detector so that they are close to the encoder disc and centered along the aperture in the disc. Secure the encoder PC board to the short ends of the blocks.

To test the circuit and align the disc, use the program shown in Listing 4. Connect the APT to your power supply and parallel port, but leave the tray off for now. Run the program and note that the encoder disc rotates. The disc should stop with the aperture directly between, or at least near, the IR devices. If it's directly under, you're done. Otherwise, rotate the disc slightly, as instructed by the program. In either case, press <Enter> to end the program. Move the shaft (not the disc) so the aperture is not under the diode and run the program again. Now, the disc should stop with its aperture directly under the diode. After completing the alignment, reinstall the tray.

Let's take a quick look at the alignment program. Lines 7–16 form a loop that continually looks for the aperture. Within that loop, lines 8–13 constitute our old stepper-motor movement

to continue experimenting with stepper motors. For example, you could modify the program to include encoder-disc sensing. You could also modify the encoder disc to incorporate twelve apertures, one for each bin position.

A more ambitious project would be to use stepper motors to provide accurate X-Y positioning. In that way you could create an automated system for drilling PC boards with ease and precision. That will be our next project; look for it soon! Ω

REFLECTION TESTER

(Continued from page 62)

ment. You have positively found an impedance mismatch in the cable. You also know exactly where it is, and how long it is.

The TV-antenna-cable problem worked out great. Can the C.R.T. help you figure out why your LAN isn't working? Let's imagine a really bad scenario. You have just installed a LAN for a small business. You are using Ethernet technology in a 10Base2 configuration. You have finished hooking up all of the hardware and installed all the necessary software, but nothing works. What do you do?

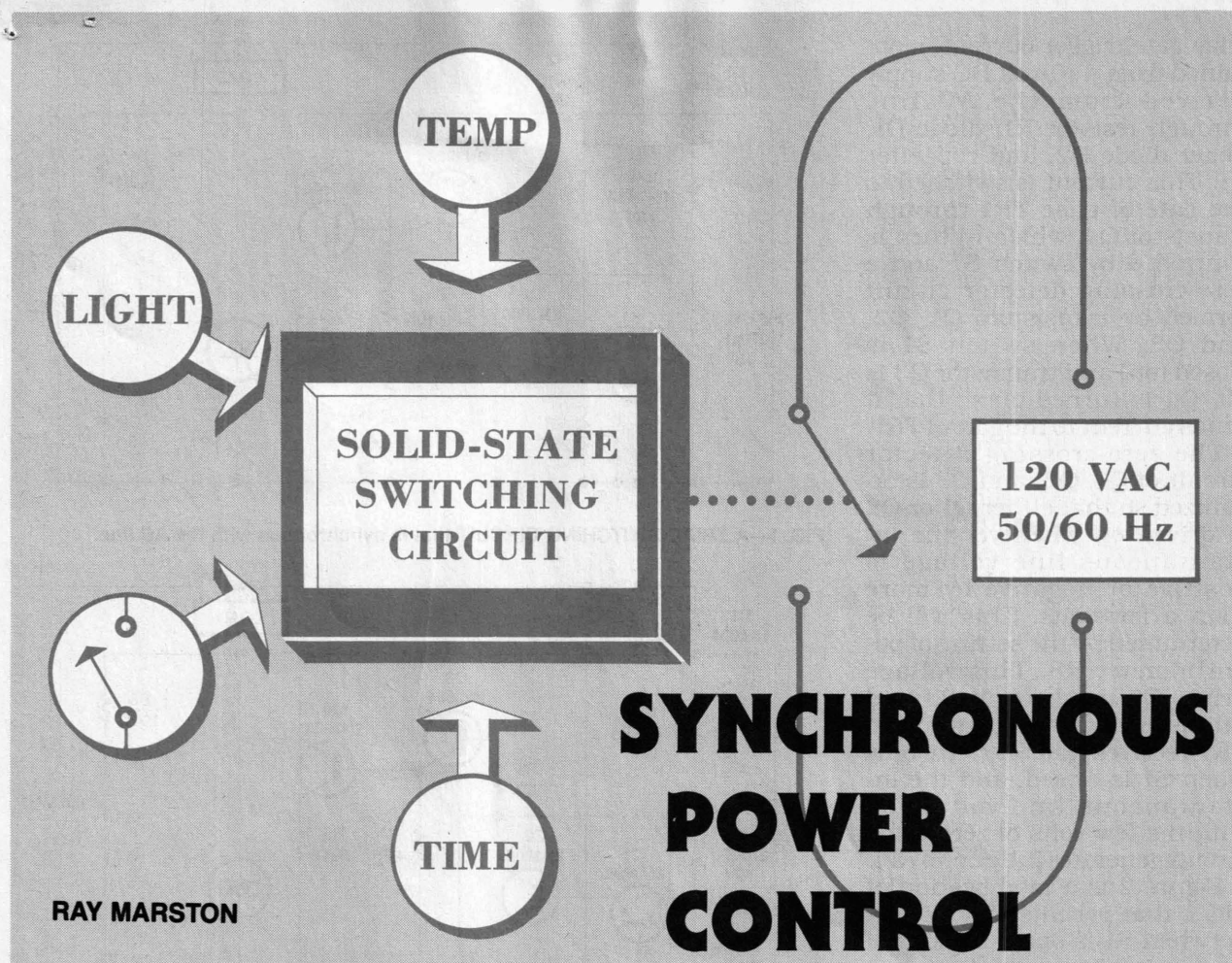
After you sit and stare for a while thinking how nice it would be camping in the middle of a stand of eastern white pine, a thought crosses your mind. Could the problem be in the cabling? You take out your C.R.T. and put it in-line with one of the workstations. The waveform on the scope looks like the display in Fig. 11. The measurement of the initial step up accounts for the 25 foot length of cable between the C.R.T. and the workstation, but at the end there is a strange step down.

A step down in the display, as we just learned in the TV antenna cable example before, indicates an impedance mismatch. But where is that step down coming from? The workstation is properly terminated with a 50-ohm terminator. But then you notice that you have wired the system with RG-59/U instead of RG-58/U. The cable is not properly matched to the Ethernet cards and terminators. The Ethernet is interpreting that mismatch as data collisions. Obviously, new cable will have to be ordered, but hopefully the cable has not yet been run under the floor tiles or through the ceiling joists.

Now that we've seen several cases of impedance mismatches, what does a healthy cable look like? Figure 12 shows a cable that is in good condition and is properly terminated with a resistor that is equal to the characteristic impedance of the line. It is a length of RG-58/U terminated with a 50-ohm resistor. If the proper cable was chosen for the LAN problem, the scope display would have looked like Fig. 12 instead of Fig. 11.

If the world were an ideal place, Fig. 12 would show a straight line after the initial ringing, but you can see that it doesn't. There is still a slight hump where the terminator connects to the cable, but it is very small and there are no multiple reflections. Also notice one other curious thing; the waveform is slowly curving down over time. That is due to the capacitive component of the line.

There you have it—a poor man's TDR. Surprising accuracy is practicable with even a crude setup. If you find the C.R.T. very useful, lay it out on printed-circuit board and mount it either in a box or in your oscilloscope. If you have any questions or problems, you can reach the author via e-mail at 75104.3104@compuserve.com. Ω



Learn about zero-voltage or synchronous AC-line power switching and control and put it to work in your own projects

THE SUBJECT OF THIS ARTICLE IS synchronous on/off power switching circuits that can switch AC power for lighting, heaters, air conditioners, motors, household appliances, and industrial and commercial equipment. Many different kinds of AC control circuits are presented including four different circuits for the control of a home electric heater. This is the third in a series of articles on power switching.

All of the circuits described here include triac drivers, and the components for all circuits were selected for operation from 120-volt, 50/60-Hz line power. The reader can select a triac with the rating and package style that best suits his applica-

tion or circuit experiment requirements.

Synchronous switching

Previous articles in this series explained that a triac is a bidirectional thyristor that can be turned on to pass a load current in either direction. It acts like two inverse-parallel-connected silicon controlled rectifiers (SCR). This characteristic permits it to function in both AC and DC circuits. Triacs can be triggered (turned on and latched) either synchronously or asynchronously when across an AC line.

A suitable triac for use in building the circuits discussed in this article would have a voltage rating of least a 200-volts

and an on-state (RMS) current rating of 4 to 6 amperes. A Motorola 2N6071 or equivalent will meet these requirements.

Recall that synchronous circuits *always* turn on at the same point in each AC half-cycle (usually just after the zero-crossing point). When triac turn-on occurs at zero crossing, little or no radio-frequency interference (RFI) is generated if the load is primarily resistive. By contrast, asynchronous circuits do not switch regularly at a fixed point on the AC sine-wave, so they can generate a significant amount of RFI.

Figure 1 is a synchronized AC power switch that is triggered near the zero-voltage crossover points of the AC waveform. The

triac gate trigger current is obtained from a 10-volt DC supply derived from the AC line through resistor R1, diode D1, Zener diode D2, and capacitor C1. This current is switched to the gate of triac TR1 through transistor Q4, which in turn is controlled by switch S1 and a zero-crossing detector circuit formed by transistors Q1, Q2, and Q3. When switch S1 is closed (on) and transistor Q3 is off, Q4 is turned on so that it sends current to the gate of TR1.

The zero-crossing detector circuit of Q1, Q2, and Q3 is organized so that either Q1 or Q2 is driven on whenever the instantaneous line voltage is positive or negative by more than a few volts. This will be determined by the setting of potentiometer R8. This voltage drives Q3 on through R3 and inhibits Q4. Consequently triac TR1 receives gate current only when S1 is closed, and the instantaneous line voltage is within a few volts of zero. This circuit generates little or no RFI.

Figure 2 is a modification of Fig. 1 that permits TR1 to turn on when S1 is open. In both of these circuits, only a narrow current pulse is sent to the gate of triac TR1. As a result, current drain on the supply is only about 1 milliamper. Switch S1 can be replaced by a combination sensor and switch that will sense changes in temperature or other physical variables, and switch the power to appliances and other AC-powered products automatically.

If the switch is replaced by an optocoupler, as will be discussed later, the control circuit will be electrically isolated from any of the external circuitry and the threat of electrical shock for the user in the event of a circuit fault will be eliminated.

Zero-voltage switching ICs

The CA3059 and the CA3079 are monolithic silicon IC zero-voltage switches capable of controlling thyristors in different applications. Originally developed by RCA, they are now made by Harris Semiconductor. These ICs are useful in AC-power switching for AC input

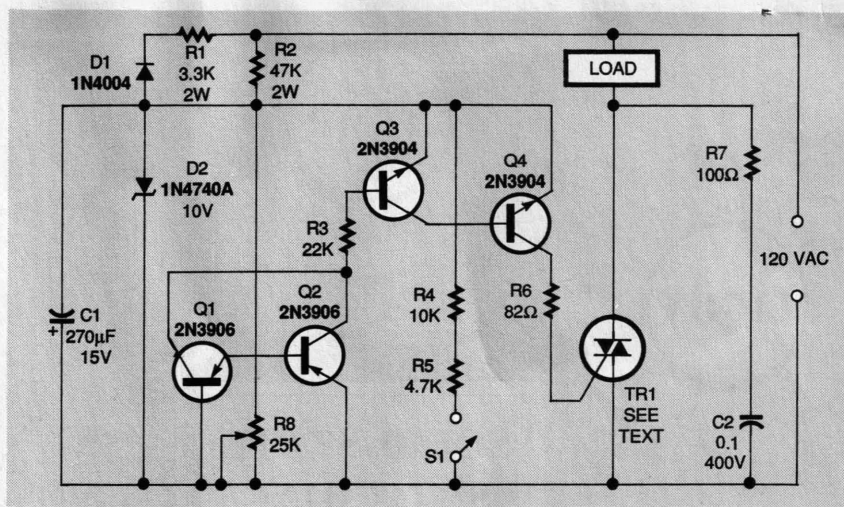


FIG. 1—A TRIAC SWITCHING CIRCUIT that is synchronous with the AC line.

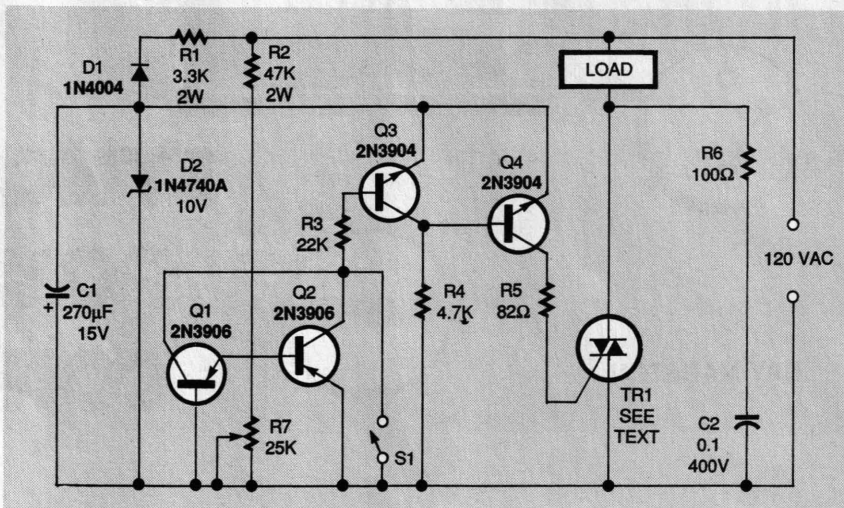


FIG. 2—A MODIFIED VERSION of the synchronous AC line switch.

voltages from 24 to 277 volts at 50/60 and 400 Hz. Each includes four functional blocks:

1. *Limiter power supply* that permits it to operate directly from the AC line.
2. *Differential on/off sensing amplifier* that tests the condition of external sensors or command signals.
3. *Zero-crossing detector* that synchronizes the output pulses of the circuit when the AC cycle is at the zero voltage point. This eliminates radio-frequency interference (RFI) when the load is resistive.
4. *Triac gating circuit* that provides high-current pulses to the gate of the power-controlling thyristor.

Figure 3 is a pinout diagram for the CA3059. It is the same as the pinout diagram for the CA3079, except that pin 14 of

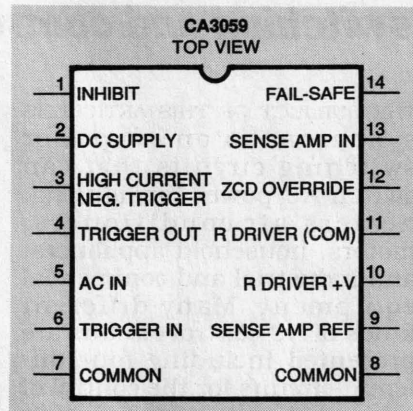


FIG. 3—PINOUT DIAGRAM for the CA3059 zero-voltage thyristor driver integrated circuit.

that IC is not to be used. The CA3079 is the same as the CA3059 except for differences in input current rating and sensor range, and it is rated for a maximum DC supply voltage of

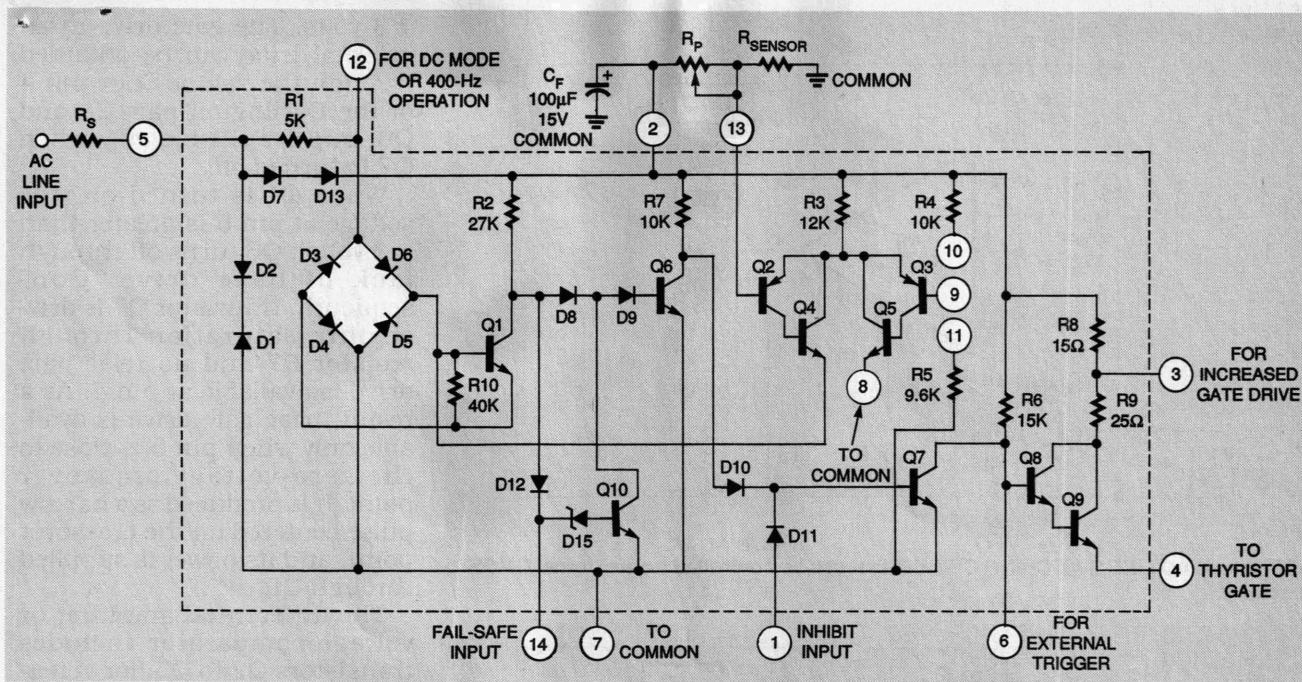


FIG. 4—INTERNAL CIRCUITRY of a CA3059 zero-voltage thyristor driver IC.

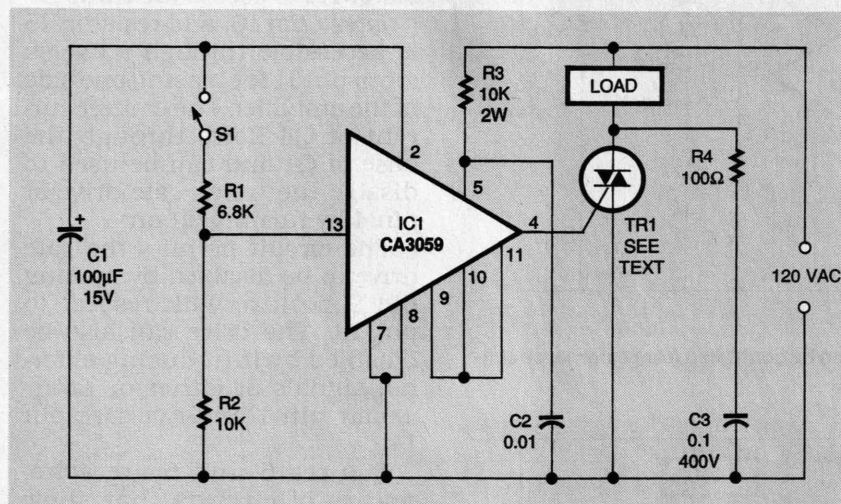


FIG. 5—A ZERO VOLTAGE TRIAC driver AC line switch based on the CA3059.

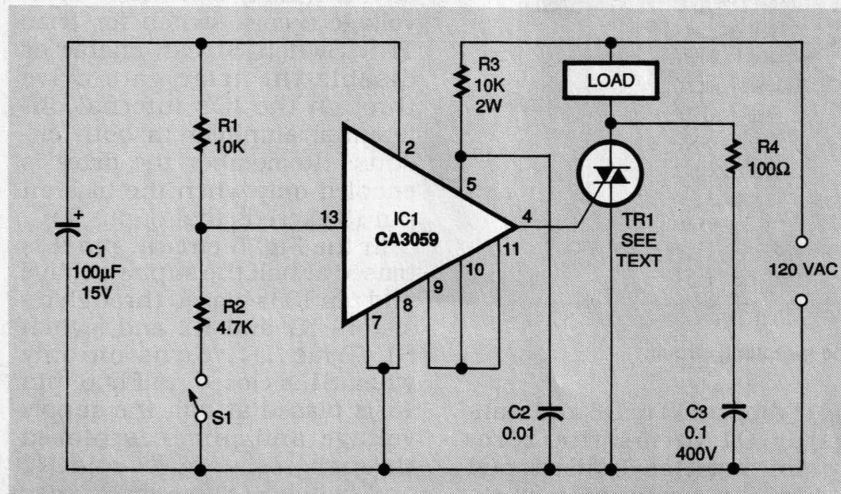


FIG. 6—AN ALTERNATIVE CIRCUIT for a CA3059-based AC line switching circuit.

10 volts rather than the CA3059's 14 volts.

The CA3059 has a built-in protection circuit that can be actuated to remove the drive from the triac if the sensor open- or short-circuits. The triac can be inhibited by the action of an internal diode gate connected to pin 1. Moreover, high-power DC comparator operation is provided by the overriding the zero-crossing detector. This is accomplished by connecting ZCD OVERRIDE pin 12 to COMMON pin 7. Gate current to the triac is continuous when SENSE AMP IN pin 13 is positive with respect to SENSE AMP REF pin 9.

Figure 4 is the internal schematic diagram for both the CA3059 and CA3079. Also included in the schematic are some important external components. The AC line is connected across AC IN pin 5 and pin 7 through limiting resistor R1. Diodes D1 and D2 act as back-to-back Zener diodes to limit the voltage on AC IN pin 5 to ± 8 volts.

On positive half cycles, diodes D7 and D13 rectify this voltage and generate 6.5 volts across the 100 μ F electrolytic capacitor C_F connected to DC SUPPLY pin 2. This capacitor stores enough energy to drive all of the IC's internal circuitry and still provide

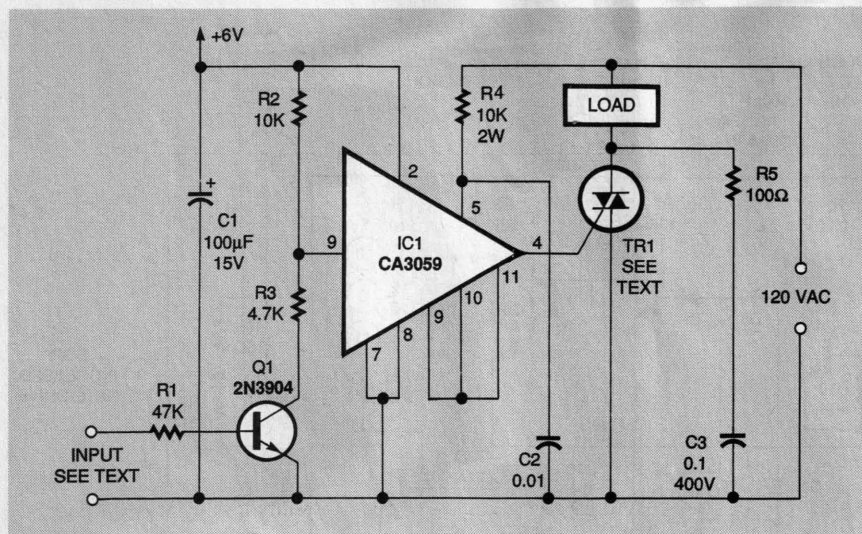


FIG. 7—AN AC LINE SWITCHING CIRCUIT that can be controlled from external CMOS circuitry.

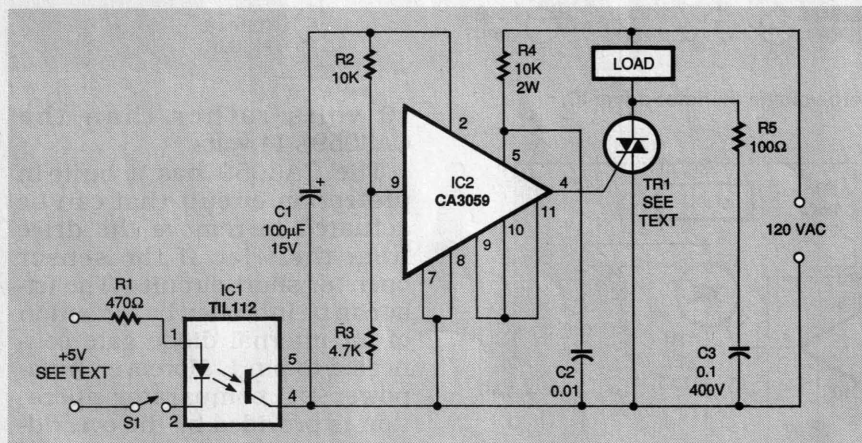


FIG. 8—AN AC LINE SWITCHING CIRCUIT that can be controlled from external circuitry through an isolating optocoupler.

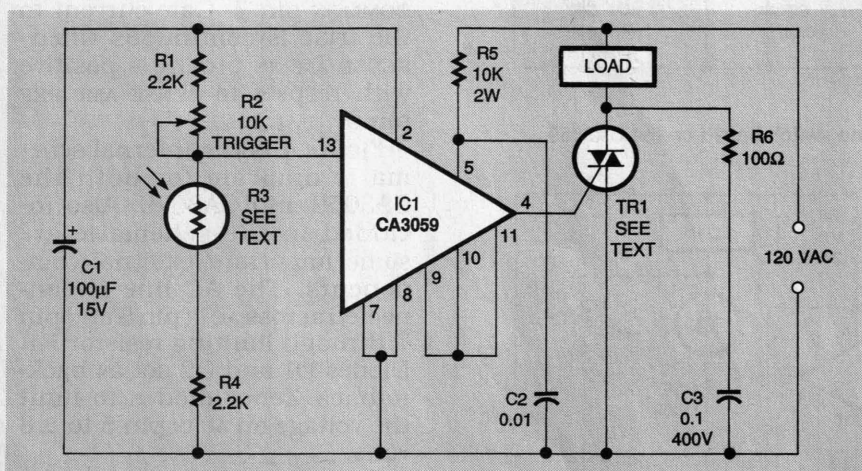


FIG. 9—A LIGHT-ACTIVATED zero-voltage AC line switching circuit.

adequate triac gate drive. A current of a few milliamperes is also available for powering external circuitry.

The bridge rectifier consist-

ing of diodes D3 to D6 and transistor Q1 forms the zero-crossing detector. Transistor Q1 is driven into saturation whenever the voltage on pin 5 exceeds

± 3 volts. The gate drive to an external triac can be obtained through the TRIGGER OUT pin 4 of the Darlington pair Q8 and Q9, but is available only when Q7 is turned off.

When Q1 is turned on (the voltage at pin 5 is greater than ± 3 volts), Q6 turns off through lack of base drive. Consequently, transistor Q7 is driven into saturation through resistor R7 and no triac gate drive is available at pin 4. As a result, triac gate drive is available only when pin 5 is close to the zero-voltage crossover point. It is produced as a narrow pulse centered on the crossover point, and its power is supplied through C1.

The differential amplifier or voltage comparator includes transistors Q2 to Q5, for general purpose applications. Resistor R4 is accessible through R DRIVER pin 10, and resistor R5 is accessible through R DRIVER (COM) pin 11 for biasing one side of the amplifier. The emitter current of Q4 flows through the base of Q1 and can be used to disable the triac's gate drive at pin 4 by turning Q1 on.

The circuit permits the gate drive to be disabled by making pin 9 positive with respect to pin 13. The drive can also be disabled by introducing external signals at either or both INHIBIT pin 1 and FAIL-SAFE pin 14.

Figures 5 and 6 are schematics of circuits that show how the CA3059 can function as a manually controlled, zero-voltage ON/OFF switch for triac TR1. Switch S1 can enable or disable the triac gate drive through the IC's internal differential amplifier in both circuits. (Remember the drive is enabled only when the bias on pin 13 exceeds that of pin 9.)

In the Fig. 5 circuit, pin 9 is biased at half the supply voltage and pin 13 is biased through resistors R1 and R2 and switch S1. Triac TR1 turns on only when S1 is closed. In Fig. 6, pin 13 is biased at half the supply voltage and pin 9 is biased through resistors R1 and R2 and switch S1. Here again, triac TR1 turns on only when S1 is

- closed. Switch S1 switches voltage with a maximum value of 6 volts and current that has a maximum value of approximately 1 milliampere.

In both of these schematics, capacitor C2 applies a slight phase delay to the AC in pin 5 of the CA3059. This shifts the gate pulse phase so that the gate pulses occur after rather than straddle the zero-voltage crossover point. The triac in Fig. 6 can be turned on by pulling R2 low or it can be turned off by letting R3 float.

Figures 7 and 8 are circuits that demonstrate how this relationship can be applied to extend the versatility of the basic circuit. In Fig. 7, the triac can be turned on and off by transistor Q1, which in turn can be controlled by external CMOS circuitry (such as one-shot or astable multivibrator). These can be powered from the 6-volt supply at DC SUPPLY pin 2 of IC1.

The circuit in Fig. 8 can be turned on and off by fully isolated external circuitry through an optocoupler. It needs an input of only three or four volts to turn the triac TR1 on. A simple optocoupler or optoisolator such as the TIL112 made by Motorola and others will work here.

Figures 9 and 10 show several of ways of organizing the CA3059 so that the triac functions as a light-sensitive (or dark-operated) power switch. In these two circuits, the CA3059's built-in precision voltage comparator turns the triac on or off when one of the comparator input voltages goes above or below the other. A stock cadmium-sulfide (CdS) photoresistor will work here.

Figure 9 is a dark (or lack of light)-activated power switch. Pin 9 of IC1 is connected to the half voltage supply, and pin 13 is controlled through resistor R1, potentiometer photoconductive cell (or photoresistor) R3, and resistor R4 acting together as a voltage divider. Under bright light the photoconductive cell has a low resistance, so the voltage on pin 13 is below that on pin 9 and the triac is disabled.

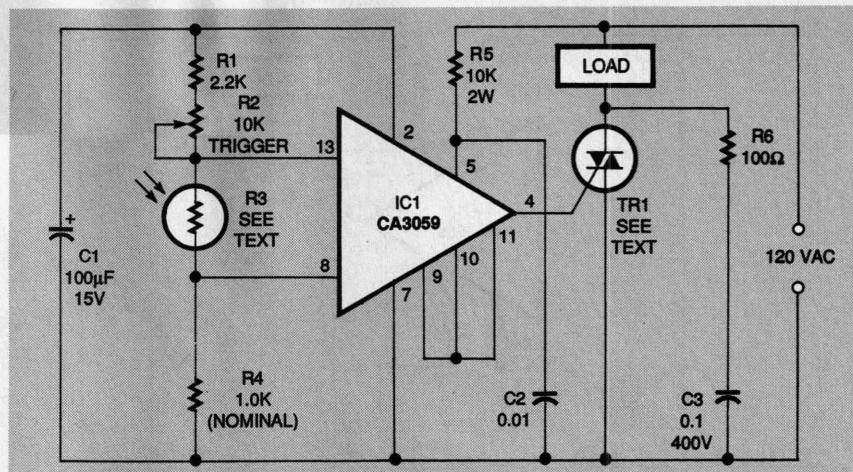


FIG. 10—A LIGHT-ACTIVATED zero-voltage AC line switching circuit with hysteresis that delays its response.

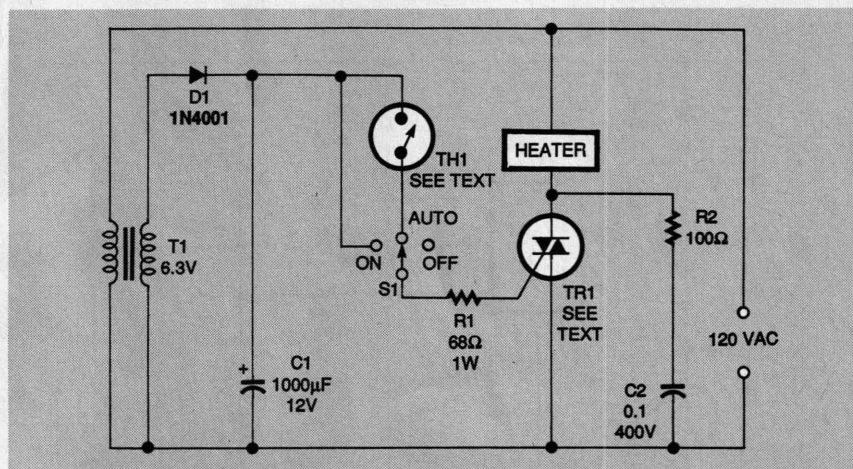


FIG. 11—AN ELECTRIC HEATER CONTROL circuit that offers thermostat-switched DC gating.

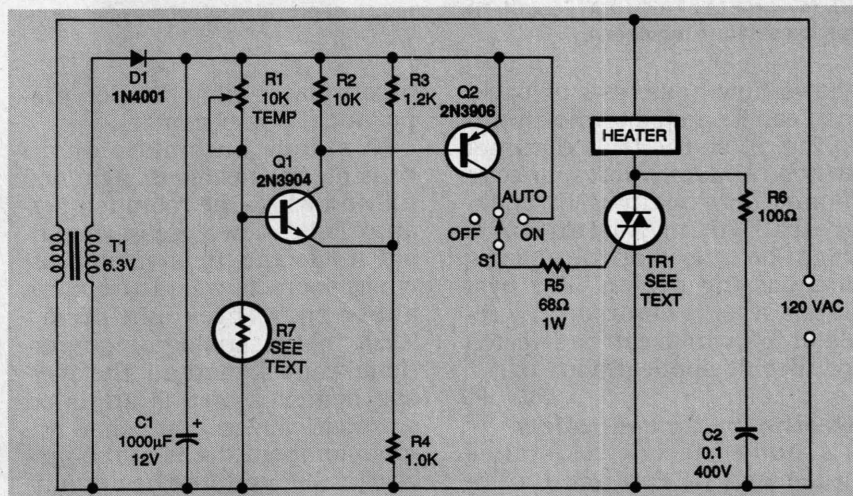


FIG. 12—AN ELECTRIC HEATER CONTROL circuit that offers thermistor-switched DC gating.

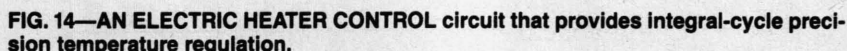
Under dark conditions the photoconductive cell has a high resistance. Therefore, pin 13 has a higher voltage than pin 9. This triggers the triac so AC

power is fed to the load. The threshold switching level of the circuit can be preset with trimmer potentiometer R2.

The schematic of Fig. 10



The advantage of synchronous gating is that no high-level RFI is generated as the triac transitions. The disadvantage is that the triac generates low level RFI continuously when in its on state.



A home electric-resistance heater can be controlled automatically to maintain the temperature of a room if the heater is switched on and off by a triac with a thermistor or thermostat acting as the feedback control element in a closed-loop system. Two different control methods are possible: simple automatic

Because of the high power requirements of electric heaters, attention must be given to minimizing radio-frequency interference in the design of triac-control circuits. This can be

By contrast, the circuit in Fig. 12 is controlled by negative temperature coefficient (NTC) thermistor TH1 and transistors Q1 and Q2. Potentiometer R1, thermistor TH1, and resistors R3 and R4 form a heat-sensitive bridge, and transistor Q1 detects bridge balance. Potentiometer R1 can be adjusted so that Q1 just starts to conduct as the room temperature falls to the preset level. Below that level, transistors Q1 and Q2 and triac

Continued on page 74

EARLY TV

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cated on one side of the scanning disk. A lens focused the image on the disk.

Another disk with a sawtooth edge was positioned between the object and scanning disk. It chopped up the light beam to a preset frequency. The object was scanned vertically, with the light-chopping disk dividing each line into a definite number of picture elements. A light-sensitive cell on the other side of the disk received the scanned image and changed the light variations into amplitude modulated current.

Baird's receiver had a similar scanning disk. Each lens in the disk projected a thin beam of light from behind the disk to a screen. As the light from each lens scanned the screen, a picture was built up. Baird synchronized his transmitter and

receiver disks by connecting an AC generator to the transmission disk's drive motor and a synchronous motor was connected to the receiving disk.

The first television drama was broadcast by station WGY in 1928, and the first television images were screen-projected in a theater in Schenectady, New York, in 1930. Both systems were mechanically scanned.

Electronic scanning

Television as we know it today became a reality only after the invention of the iconoscope or camera tube by Vladimir Zworykin, a Russian emigre to the U.S. The iconoscope, when paired with the receiving picture tube, made electronic scanning possible. Both the iconoscope and the picture tube were developed from the earlier invention of the cathode-ray tube.

With the development of the camera tube and receiving tube, electronic scanning was able to

replace cumbersome and expensive mechanical scanning. The image orthicon, invented by Philo Farnsworth in 1928, improved upon Zworykin's iconoscope, but it was not until 1933 that electronically scanned television became a commercial reality.

Who named television?

Two men claimed to have coined the word television. One was Hugo Gernsback, then the publisher of *Radio News* (He later founded Gernsback Inc., the publisher of *Radio-Electronics* which was renamed *Electronics Now* in 1993).

In the August 1928 issue of *Radio News*, Gernsback wrote: "The word 'television' was first coined by myself in an article entitled 'Television and the Telephot' which appeared in the December 1909 issue of *Modern Electrics*." The other person who claimed to have coined the word in 1900 was a Frenchman named Perskyi. Ω

POWER CONTROL

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TR1 are all driven full on. Above that preset temperature threshold, all three components are cut off.

The gate-drive polarity in Fig. 12 is always positive, but the triac's main-terminal current is alternating. As a result, the triac is gated alternately. A triac's gate sensitivity differs, depending on its operating mode. Thus, when the temperature sensed by the thermistor is well below the preset temperature threshold, Q2 is driven full on and the triac is gated in two quadrants. This sends the full AC-line power to the heater.

However, when the room temperature is very close to its preset value, Q2 is driven only slightly on. Thus triac TR1 is gated on in only one mode and only half of the available AC-line power is available to the heater. Consequently, the circuit offers fine temperature control of the room.

Synchronous switching

Figure 13 is the schematic for an electric heater-control circuit with automatic zero-voltage synchronous AC-line switching based on the CA3059 (IC1). It provides simple on/off heater switching.

This circuit is similar to that of the light-activated power switch of Fig. 9 except that an NTC thermistor is the feedback sensor. The Fig. 13 circuit is capable of maintaining room temperature within one or two degrees Fahrenheit of the value set on potentiometer R1.

Figure 14 is the schematic for a proportional heater control circuit that is capable of regulating room temperatures to within one degree Fahrenheit of a preset value. A thermistor-controlled voltage is applied to pin 13 of IC2 (CA3059). Simultaneously, a repetitive 300 millisecond ramp waveform from pins 2 and 6 of IC1 (a 7555 or CMOS 555) is applied to pin 9 of IC2. The 7555 is configured as an astable multivibrator, and the waveform is centered on the half-supply voltage.

The triac is turned fully on synchronously if the room temperature is more than a few degrees below its preset level. However, it is cut off fully if the room temperature is more than several degrees above that preset level. When the temperature is within a few degrees of the preset value, however, the ramp waveform synchronously turns the triac on and off (in the integral cycle mode) once every 300 milliseconds with a duty cycle that is proportional to the temperature differential.

For example, if the duty cycle ratio is 1:1, the heater generates only half of its maximum power. However, if the ratio is 1:3, it generates only one quarter of its maximum power. This response causes heater output power to self-adjust to meet the room's preset heating demand.

When room temperature reaches the precise preset value, the heater does not switch fully off. Rather, it generates just enough output power to compensate for the room heat loss. This system gives precise room temperature control. Ω

William Sheets, K2MQJ and Rudolf F. Graf, KA2CWL

**Pull in ATV transmissions
on the 900 MHz and
1300 MHz bands**

BUILD THIS UHF DOWNCONVERTER

THE MINIATURE, TUNEABLE UHF downconverter described in this article is designed to receive amateur television (ATV) signals in the 900–1300-MHz band and convert them for display on an ordinary TV set. The downconverter is the perfect match for the 900-MHz ATV transmitter described in last month's issue of *Electronics Now*. It is also used to receive signals from low-powered video links operating in the 900 (33-centimeter) band.

As a bonus, a design for 1300 MHz downconverter is also presented to allow reception of any ATV activity in the 23-centimeter (1240–1300 MHz) amateur band. Both the 900-MHz and 1300-MHz converters are mechanically nearly identical and appear the same unless inspected closely. They differ slightly with regard to the PC board microstrip tuned elements and minor difference in the local oscillator circuit. Otherwise the same circuit architecture and parts are used for both. Their theory of operation and construction are closely

paralleled. In the discussion that follows, reference is made only to the 900-MHz unit unless differences between the two units are noted.

The downconverter is not intended for reception of narrow-band FM or SSB signals since the converter uses a tuneable local oscillator covering a 30 to 60 MHz range. The 900-MHz and 1300-MHz versions are useable for wide-band modes such as AMTV, FMTV or FM audio reception using a conventional FM broadcast receiver tuned to 90 MHz or so on the dial.

Basic signal flow

The downconverter block diagram is shown in Fig. 1. Signals from the RF input are fed to low-noise RF amplifier Q1, which boosts the antenna signal and feeds it to mixer transistor Q2. There the amplified signal is mixed with the output of the tuneable voltage-controlled local oscillator Q5 to an intermediate frequency (IF) of 60 MHz. The IF signal is fed from the mixer to an IF pre-amplifier stage.

Three tuned circuits, Z1–Z4, are used in the RF amplifier stage. These are not discrete components—they are of PC board stripline construction, and consist only of circuit-board traces. Other tuned stripline circuits, Z5–Z6 are also used in the downconverter.

Voltage-controlled oscillator Q5 is tuned by a DC voltage. This DC voltage can be selected from either a PC board potentiometer or via an external, coaxial-fed power supply that allows the downconverter to be tuned remotely. This arrangement enables the downconverter to be mounted on the right at the antenna feed point to eliminate feed-line losses that are significant at this frequency range. No separate DC feed is necessary since the coax (RG59/U or RG6/U recommended) carries both DC power, tuning voltage and IF signal.

The downconverter's input impedance is 50 ohms, and any matching 900-MHz or 1300-MHz antenna can be used. The overall gain of the downconverter is about 37–43 dB. The

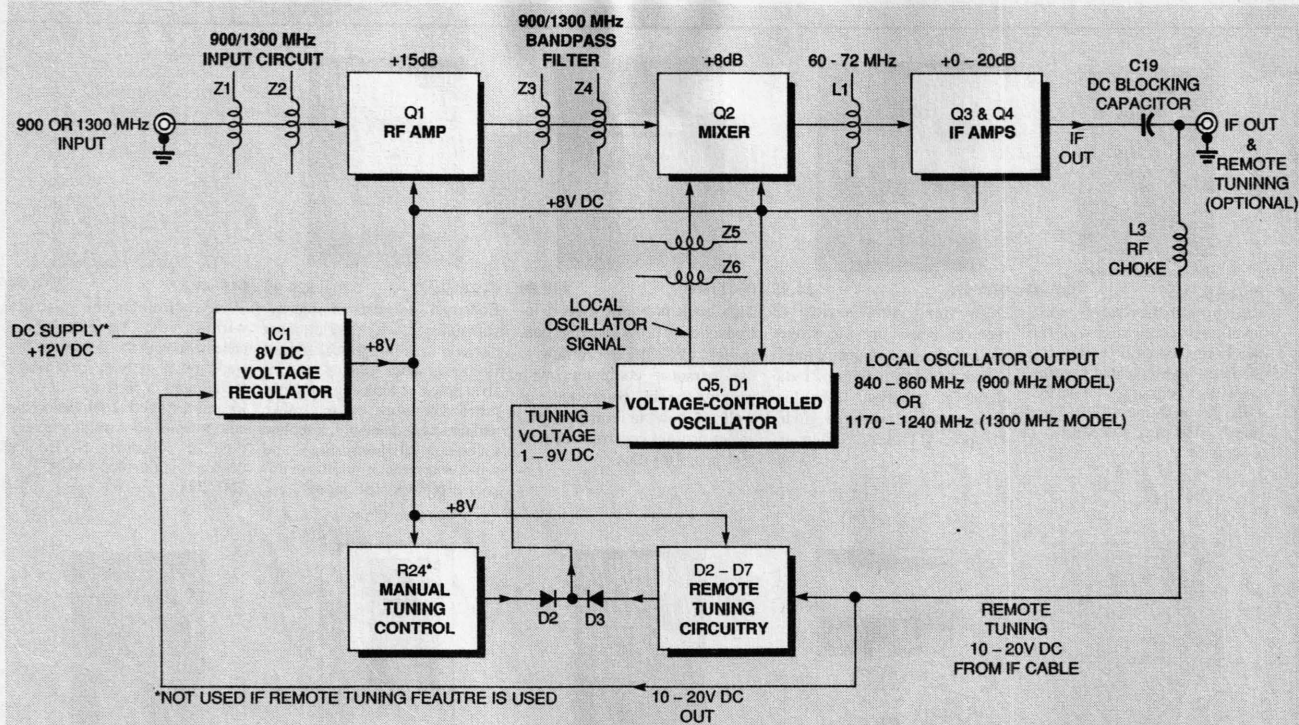


FIG. 1—BLOCK DIAGRAM for the downconverter serves for both the 900 and 1300-MHz versions.

gain can be changed as needed by changing a resistor value.

For portable operation or for use in a packaged ATV transceiver or a video handy talky (HT), the on-board tuning potentiometer may be used. The tuning voltage is tapped from the +8-volt DC voltage-regulator, IC1.

For on-board tuning, the external power supply must deliver a fixed 11 to 20 volts DC at approximately 30 milliamperes to the 8-volt DC regulator. For remote tuning, the supply must be variable.

Circuit operation

The downconverter schematic is shown in Fig. 2. The input signal from antenna jack J1 is fed to a filter made up of Z1, C1, and Z2. Microstrip lines Z1 and Z2 are part of the PC board foil surface and form a tuning circuit. (All six microstrip lines are given Z designations.) Trimmer capacitor C1 is adjusted for best reception of a weak signal. The signal is then fed to gate G1 of Q1. Transistor Q1 is a low-noise MESFET with a noise figure about 1 dB. Unavoidable circuit losses will raise the actual noise figure to typically 1.5 to 2

dB, which is fairly good for the intended application. The source S of Q1 is biased by R1. The other gate, G2, is biased by R2 and R3. Two pairs of 100-pF capacitors (C2 through C5) bypass both the source and G2 of Q1 to ground, providing the necessary bypassing for inductance Z3. Chip capacitors were chosen for C2-C5 because conventional 100-pF disc capacitors function poorly at 900 or 1300 MHz.

The amplified signal at the drain of Q1 is fed to a double-tuned coupling network consisting of Z3, C8, Z4, and C9. Trimmer capacitors C8 and C9 determine the bandpass characteristics of the network. Capacitors C6 and C7 provide RF grounding for the low end of Z4. The spacing of Z3 and Z4 on the PC board determines the amount of RF mutual coupling. This amount is fixed by the PC etched layout. The RF gain of the stage is about 12 to 15 dB. This gain figure is a net number that includes losses in the input and output tuned-circuit networks, and between the antenna input and the mixer input.

Transistor Q2 serves as a RF mixer. The bipolar transistor is

biased by R5, R6, R7, R8, and a chip resistor network that consists of R14, R15 and R16. The mixer's emitter current is typically 1 milliamper.

The amplified RF signal from Q1 is fed via capacitor C10 to the base of Q2. The local oscillator's output signal is fed from Q5 to the emitter of Q2 via microstrip feedline Z5, R14, R15, and R16. Capacitor C13 provides a low impedance at the IF frequency (61.25 or 66.25 MHz) for the emitter of Q2. The local oscillator frequency is equal to the received RF signal *minus* the IF (low side). The signal gain from the mixer circuit is typically 6 to 8 dB. Overall gain so far is typically 20 dB.

The IF signal from the mixer is fed to a matching network consisting of C11, C12, L1, L2, and C14. The matching network rejects mixer-output signals above 100 MHz. The desired signal is then fed to an IF amplifier stage consisting of Q3 and Q4. Resistor R10 provides bias voltage while R9 provides termination for the matching network. Capacitor C16 blocks DC to the base of Q3. Current gain for the matching network is set by R11 and R12. Overall IF gain is about 20 dB.

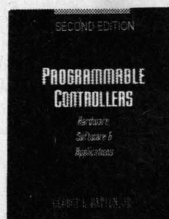
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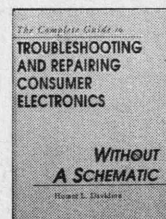
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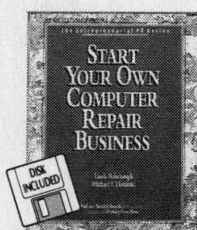
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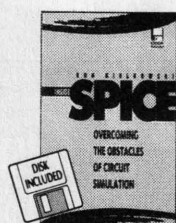
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ATV DOWNCONVERTER

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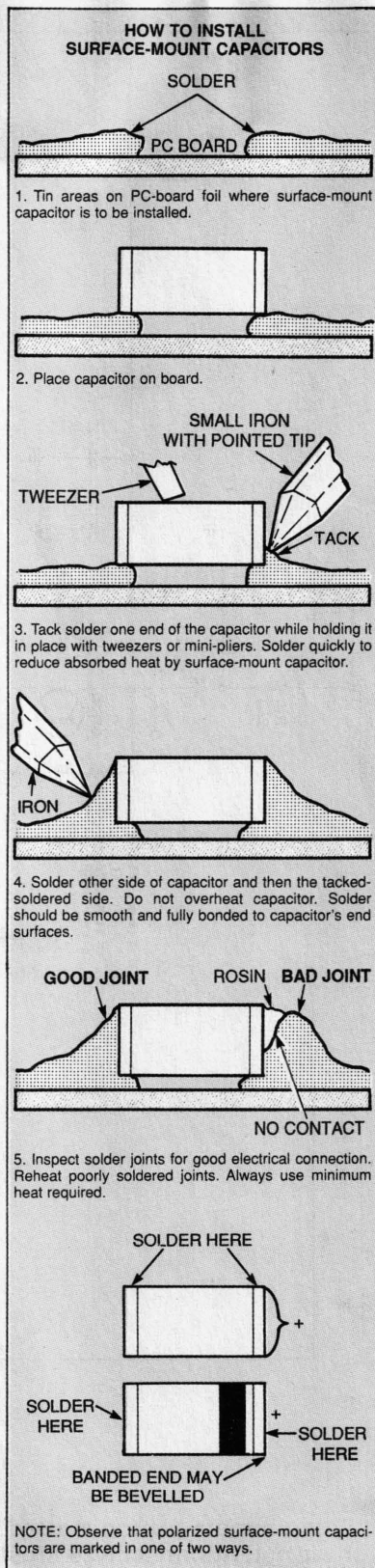
The gain for the overall circuit can be adjusted by changing the resistive value for R11 (see Table 1).

The IF amplifier directly feeds a 50 to 75-ohm load. Maximum downconverter IF output is around -2 dBm for 1 dB compression. This is 150 to 200 millivolts into a load of 50 to 75 ohms. Capacitor C18 and R13 provide decoupling of the IF circuits from the regulated 8-volt bus.

The local oscillator output signal is generated by Q5 in a typical Colpitts oscillator circuit. Bias for Q5 is provided by R17 through R21. Resistor R21 and surface-mount bypass capacitors C24 and C25 decouple the local oscillator frequency from the regulated 8-volt DC bus. The resonant circuit for the local oscillator is formed by C21, the internal base-emitter capacitance of Q5, and emitter-to-ground capacitance, strip-line Z6, trimmer capacitor C22, and varactor diode D1 in parallel with C23. (Capacitor C23 is not used in 1300 MHz version.) The capacitance of varactor D1 is varied by a voltage applied to it via R22 that ranges from 1 to 8-volts DC.

The varactor D1 provides fine tuning of the local oscillator frequency because the setting of trimmer C22 has the major effect on oscillator frequency. Trimmer C22 will change the oscillator frequency by several hundred MHz. When set correctly, C22 sets the local oscillator frequency so that D1 can tune the circuit over the desired range. This will be a bandwidth of about 30 MHz for the 900 MHz downconverter, and 60 MHz for the 1300-MHz version.

Voltage for D1 is fed through isolating resistor R22 from either of two sources. The first source is the on-board tuning potentiometer R24, whose wiper provides 1 to 8 volts DC. Resistor R25, connected to the ground side of potentiometer R24, sets the low-voltage limit to about 1 volt. Diode D2 acts as



half of a logical OR gate, and R23 and C26 provide a pull-down to ground and RF bypassing for the varactor tuning voltage. The second source is a tuning volt-

age derived from a remote-tuning setup.

Remote tuning

Remote tuning is an important feature of the downconverter because it allows the unit to be mounted directly at the antenna feed point, thus eliminating a coaxial feedline for 900-MHz signals. Feedlines have high losses at this frequency unless they are special, expensive types. Even then, losses are still high for long cable runs. For example, a good quality, expensive (\$60 to \$80 per 100 feet) coax feed-line such as 9913 has a loss of 5-dB per 100 feet of cable at 1000 MHz. The output of the downconverter can be fed via an inexpensive coaxial cable (such as the RG59 or RG66), which has an attenuation of about 2 dB per 100 feet at the output frequency and costs about \$15 per 100 feet.

A DC Block Unit (see Fig. 3) allows a variable tuning voltage to be added into the IF coaxial feed-line to power and tune the downconverter. This voltage should vary from 10 to 20 volts DC to properly remote tune the downconverter. Capacitor C19 (Fig. 2) blocks this DC but passes the IF signal from Q4. Inductor L3 passes DC (only 10-ohms DC resistance) but blocks the IF signal because it has a very high impedance at 60 MHz. The input of 10 to 20-volts DC is passed to voltage regulator IC1 via blocking diode D7 (Fig. 2). Voltage regulator IC1 supplies +8-volts DC to the downconverter circuitry, insensitive to the 10 to 20-volt variation of the input DC voltage.

The 10 to 20-volts DC input is also supplied to Zener diode D6 and diode D5, dropping 9 volts across these diodes, resulting in 1 to 11-volts DC across pull-down resistor R27. Resistor R6 and clamp diode D4 limit the voltage at the input of D3 to 1 to 9 volts. This results in about 0.3 to 8.3 volts DC across R23 and C26. This voltage is fed to varactor D1 via R22 for tuning. As the DC voltage on the IF cable varies from 10 to 20 volts, the converter is both powered and tuned over its frequency range. If oper-

ation from a fixed DC power supply (11 to 15 V) is desired, the converter is powered with 12-volts DC via the DC input and blocking diode D8. In this case, tuning is done with the on-board tuning potentiometer, R24, and the DC Block unit is not needed.

When remote tuning is used, tuning potentiometer R24 must be set at its extreme counterclockwise position to set the voltage at the wiper to its lowest DC potential. All or part of the downconverter tuning range won't be obtained with R24 in any other position.

Clean power supply

Any hum or noise on the tuning voltage from an external power supply will modulate the local-oscillator frequency, thereby frequency modulating the IF signal. This effect is most degrading for the 1300-MHz model, where the tuning sen-

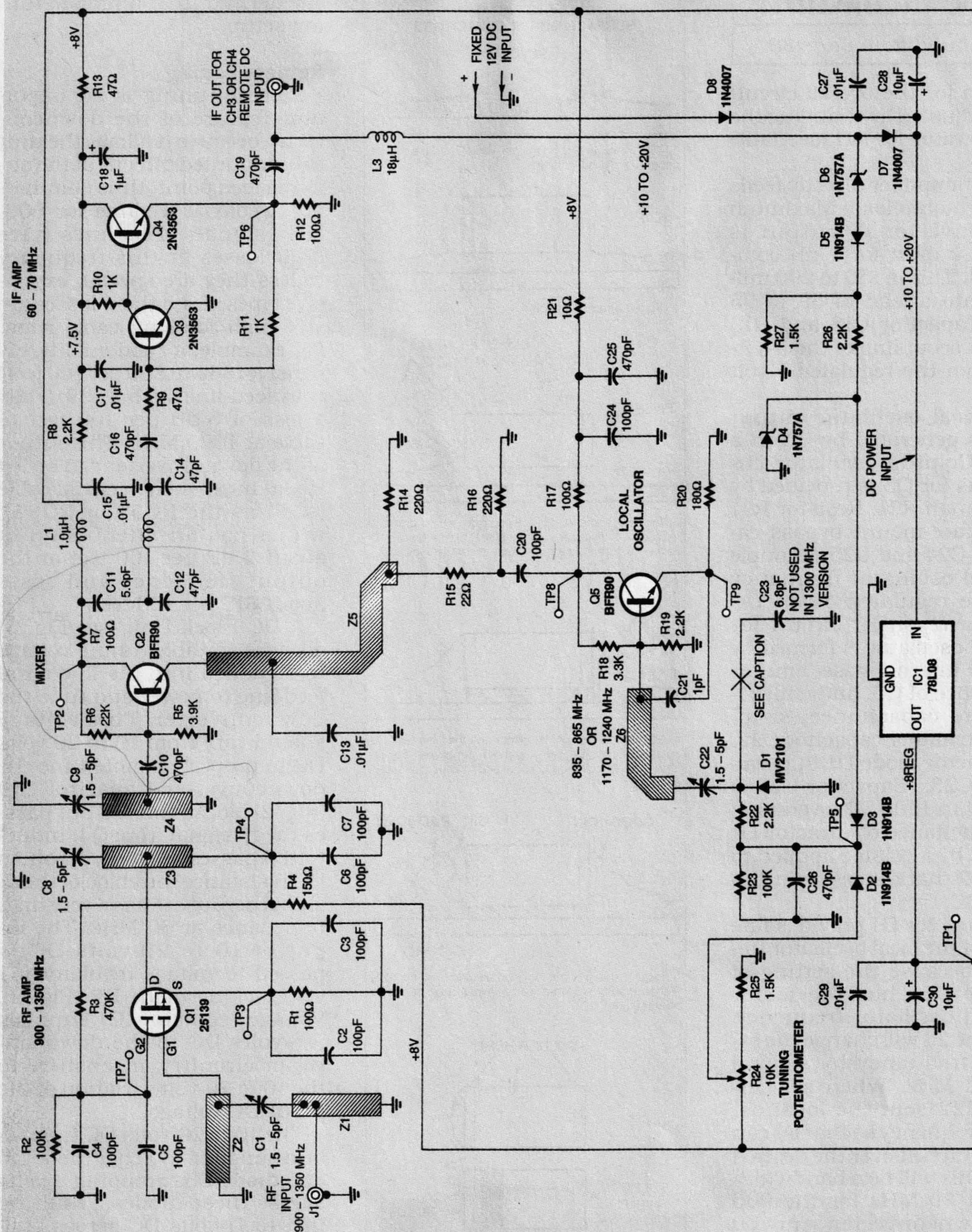


FIG. 2—SCHEMATIC DIAGRAM for the downconverter covers both the 900-MHz and 1300-MHz versions. Aside from slight foil modifications, the only difference is the elimination of the 1N914B diode in the 1300-MHz version.

TABLE 1—TROUBLESHOOTING VOLTAGES AT TEST POINTS

Test Point Location	Volts DC	Localized Trouble Area
Junction of D8, IC2	+11.4	D8
Junction of IC1, C30 (TP1)	+7.6 to 8.4	IC1
Junction of R4, C6, C7, Z2 (TP4)	+6.2 to +7.2	Q1, R4, C6, C7
Junction of R1, C2, C3	(TP3)	+0.5 to +1.2 Q1, R1, C2, C3
Junction of R2, R3, C3, C4 (TP7)	+1.0 to +1.4	Q1, R2, R3, C4, C5
Collector Q2 (TP2)	+5.5 to 7.2	Q2, R5, R6, R7, R8, R13, R14, R15, R16, L1, C11, C15, C17, C18
Junction of D2, D3, C26 (TP5)	0.5 to >+7.0 as R24 is rotated	D1, R24, R25, D2, D3, R23, R22
Collector Q5 (TP8)	+6.5 to +7.2	R21, M C24, C25, Q5, C20, R18, R19, R20
Emitter Q4	+0.8 to +1.0	Q3, R10, R11, R12, C16, C19
Base Q4	0.7 volts more than base of Q3	Q3, R10, R11, R12, C16, C19

*Check all transistors and IC's for correct installation. Check diodes and electrolytic capacitors for correct polarized installation.

sitivity of the local oscillator is 10 MHz per volt of tuning voltage. Just 1 mV of hum will result in 10 kHz of frequency deviation. For FMTV use, this will be noticeable as video hum bars in the received picture. AM signals will be less susceptible to hum and noise from the power supply.

This power supply defect should not be tolerated because it is easy to build a *clean* external power supply from inexpensive readily available voltage regulator ICs.

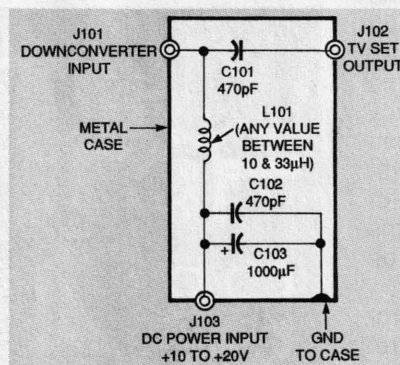


FIG. 3—DC BLOCK UNIT is required at the TV receiver or monitor site when remote tuning the downconverter.

Pre-construction details

Do not substitute parts specified in the Parts List or change the PC-board layout should you etch a PC board from the patterns shown.

Use good lighting while assembling parts on both sides of the 2.5 × 4-inch PC board. A magnifier lens is necessary because the parts are too small to check for markings and solder joints are too minuscule to inspect with the naked eye.

All trimmer capacitors and grounded leads of resistors

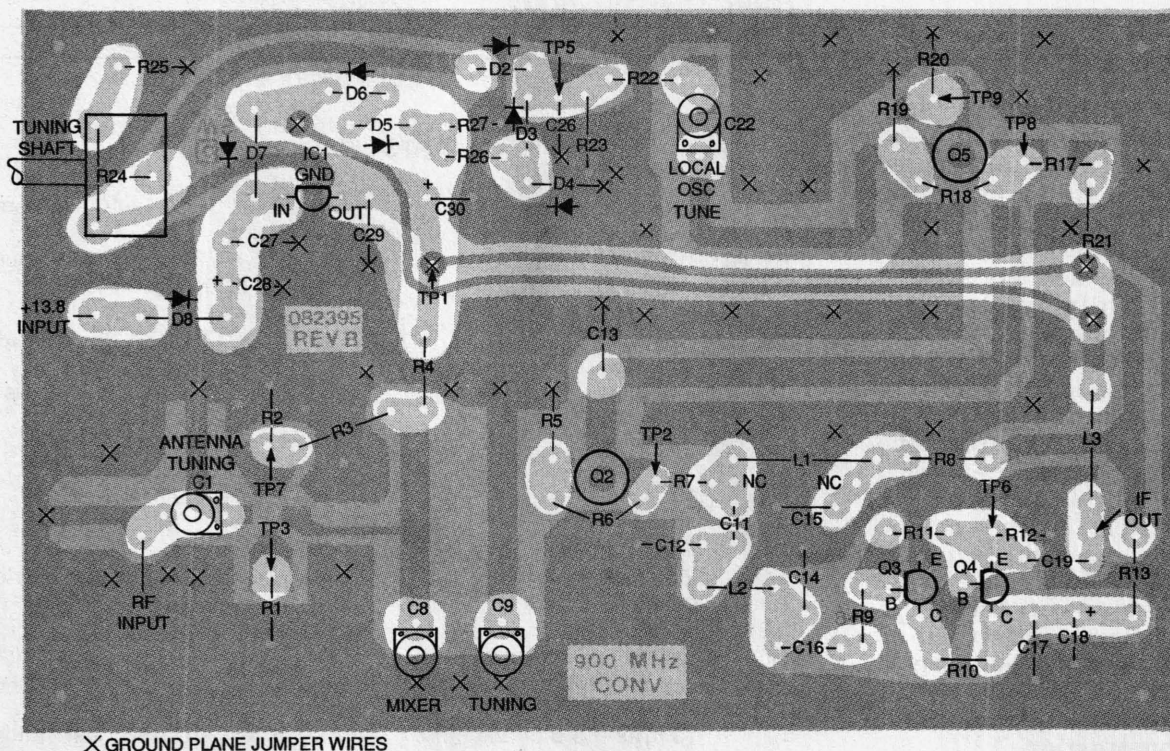
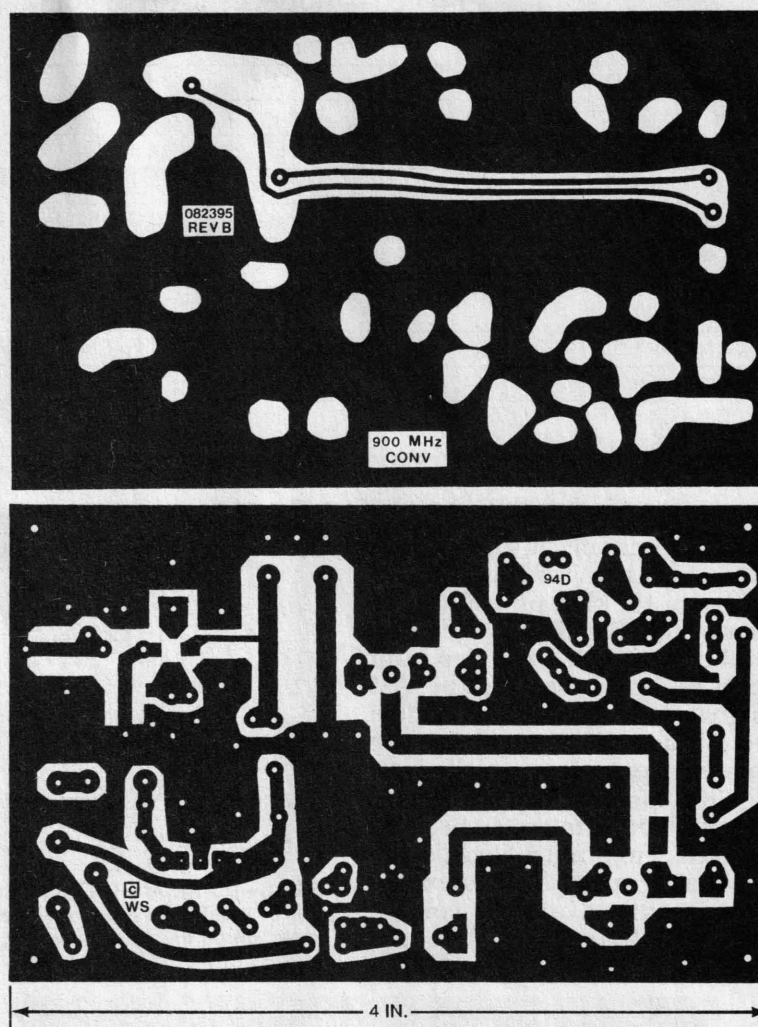


FIG. 4—TOP SIDE OF PC BOARD for the 900-MHz downconverter is shown with parts locations identified. Be sure to install all ground jumpers where indicated and solder to both sides of the PC board.

FOIL DIAGRAMS FOR THE DOWNCONVERTER 900-MHz AND 1300-MHz MODELS

While it is possible to design the same printed-circuit board foil pattern for both Downconverter models (since the frequencies differ by 350 MHz), compromises are necessary in the RF microstrip elements which would result in performance sacrifice in RF gain and tuning rate. Therefore, individual foil patterns were designed for each model. The 900-MHz model can be re-tuned to cover up to about 1100 MHz, and the 1300-MHz model will cover about 1150 to 1450-MHz (in 60-MHz segments) for other applications, with almost the same performance. The IF output frequency can be changed to anywhere between 40 and 100 MHz by changing one component and shifting the local oscillator frequency as required. The foil patterns provided are actual size and can be copied.

Starting in the upper-left corner and proceeding clockwise, the foil patterns are: 900-MHz top, 1300-MHz top, 1300-MHz bottom, and 900-MHz bottom. The PC boards are all the same size—4 X 2.5 inches.



must be soldered on both sides of the foils on the PC board. This is good RF grounding practice. Also, there are many ground-plane jumpers (approximately 3 dozen) that must be soldered in place, because the commercially available PC boards are not plated through, and plated-through PC boards cannot be made at home.

Do not solder any parts to the PC board until as many components as possible are inserted. Coils L1 through L3 are installed after all other components are inserted. Mount all parts tight and close to the board. Projects in the microwave frequency region require *zero-lead lengths* for parts in the RF circuit sections.

Start construction

Begin construction by inserting all trimmer capacitors in

PARTS LIST

All resistors are 1/8-watt, 5% unless otherwise specified

R1, R7, R12, R17—100 ohms
R2, R23—100,000 ohms
R3—470,000 ohms
R4—150 ohms
R5—3,900 ohms
R6—22,000 ohms
R8, R19, R22, R26—2,200 ohms
R9, R13—47 ohms
R10, R11—1000 ohms
R14, R16—220 ohms, surface-mount
R15—2 ohms, surface-mount
R18—3,300 ohms
R20—180 ohms
R21—10 ohms
R24—10,000-ohms potentiometer
R25, R27—1,500 ohms

Capacitors

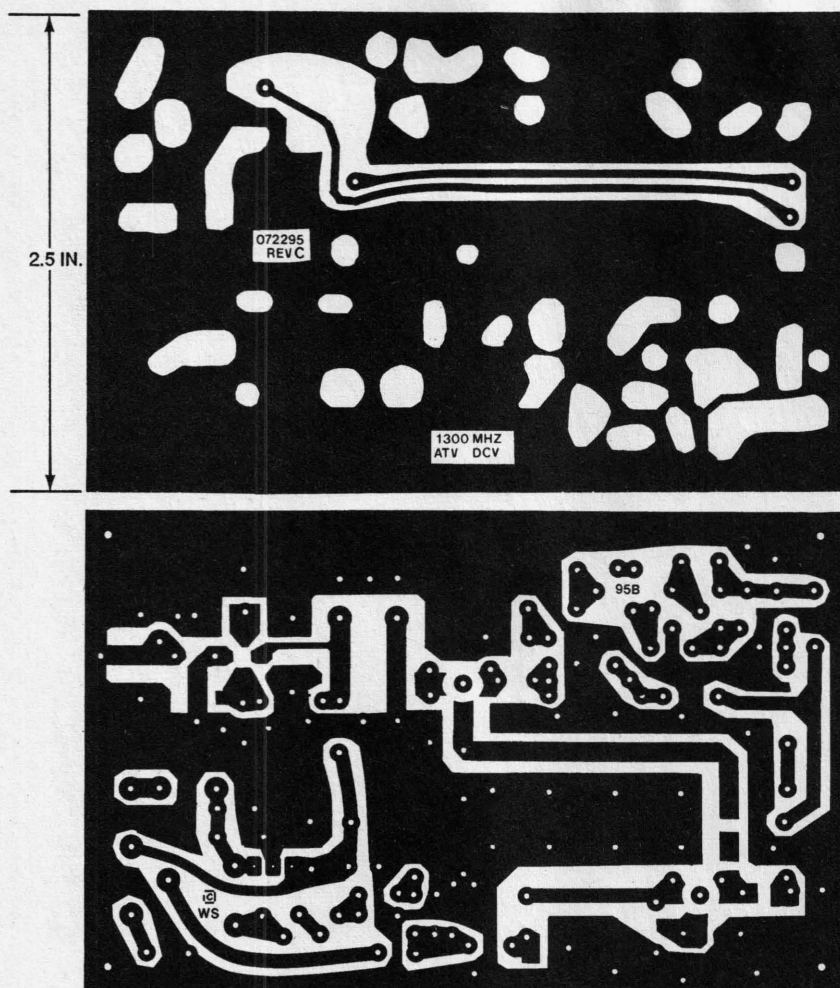
C1, C8, C9, C22—1.5-5 pF trimmer, PC mount

C2—C7, C20, C24—100 pF, surface-mount
C10, C25—470 pF, surface-mount
C11—5.6 pF, NPO
C12, C14—47 pF, NPO
C13, C15, C17, C27, C29—0.01 μ F, GMV disk, 50-volt
C16, C19, C26—470 pF, GMV disk, 50-volts
C18—1.0 μ F, 35-volt electrolytic
C21—1 pF, surface-mount
*C23—6.8 pF, surface-mount
C28, C30—10 μ F, 25-volt, electrolytic

*Used in 900 MHz downconverter only

Semiconductors

D1—MMBV2101T1 (Motorola) varactor, 6.8 pF at 4 volts, surface-mount
D2, D3, D5—1N914B, fast-switching diode



D4, D6—1N757A, 9.1-V, 0.5-watt, Zener diode
 D7, D8—1N4007, 1000-V, 2.5 A, rectifier
 Q1—NE25139 (California Eastern Labs.) low-noise, dual-gate RF-amplifier
 Q2, Q5—BFR90, npn, low-noise, UHF amplifier, MESFET
 Q3, Q4—2N3563, npn, RF/IF/Video bipolar
 IC1—78L08, positive voltage regulator

Coils and Chokes

L1—RF choke, 1 μ H, $\pm 5\%$
 L2—Fabricated coil (see text)
 L3—RF choke, 18 μ H with less than 10-ohms resistance

Miscellaneous

J1, J2—See text
 1—Plastic shaft extension for R24
 1—Length of #22 enameled, solid copper wire

1—Machine screw, #8-32, binding head

1—PC board for 900-MHz or 1300-MHz downconverter

DC BLOCK UNIT (OPTIONAL)

C101, C102—470 pF, disc ceramic, 50-volts

C103—1000 μ F, 25-volts, electrolytic

J101, J102—Type F connectors, female, chassis mount

J103—DC power connector that can be shielded, 2.5 mm or RCA phono, female chassis mount

L101—Any value from 10 to 33 μ H, less than 10-ohms DC resistance

1—Metal enclosure, RF tight

A complete parts kit for the 900-MHz or 1300-MHz downconverter, consisting of the PC board and all parts that mount on it, is available from: North Country Radio, PO Box

the PC board. See Fig. 4. Solder trimmer rotor terminals to the top side of the PC board. Be careful not to melt the plastic. Do not solder the bottom foil at this time.

Mount all $\frac{1}{8}$ -watt resistors. Only solder the top of the PC board where resistor lead passes through the ground foil. Solder as many as you can. Do not solder the bottom foil yet. Surface-mount resistors will be installed later. Install tuning potentiometer R24. Leave the thumbwheel of R24 in the center-rotation position.

Install the fixed capacitors. If any come with pre-formed leads, straighten the leads with pliers so the capacitor can fit as close to PC board as possible. Pull the leads through board so capacitor bodies are tight to PC board. Watch polarity of all electrolytic capacitors. Confirm that the capacitor values are correct prior to soldering them in place.

Do not install any of the surface-mount components yet. Install all transistors and diodes, except Q1 (25139) which will be installed later. See Figs. 4 and 5. Check polarized parts for correct orientation. All TO-92 transistors should be $\frac{1}{8}$ inch from

53, Wykagyl Station, New Rochelle, NY 10804. The 900-MHz model sells for \$39.00, 1300-MHz sells for \$41.00. Both models: \$76.50 (must be ordered at same time). Please add \$4.50 for postage and handling for one unit or \$5.50 for both units. New York residents must include local NY sales tax. Be sure to specify which model (900 or 1300 MHz) you are ordering. Parts for optional DC Block Unit are not included in this kit.

A catalog containing information on transmitters, downconverters, video cameras and lenses, and a number of other kits and electronic projects is available from the above address. Please send SASE (75-cents postage) and \$1.00 to cover handling and mailing (refundable with order).

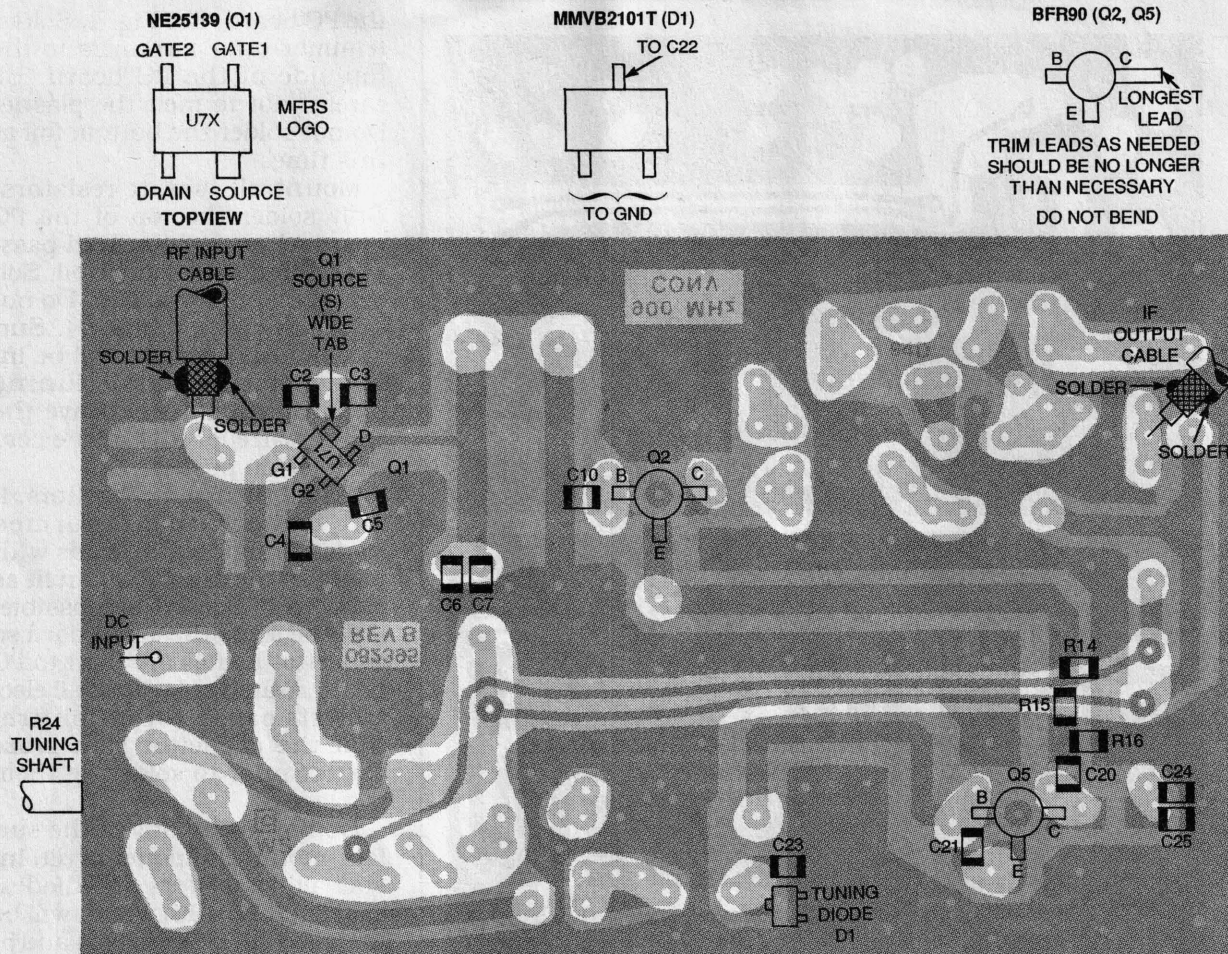


FIG. 5—BOTTOM SIDE OF PC BOARD for the 900-MHz downconverter is shown with parts locations identified. Cable lengths should be just long enough to provide suitable length when PC board is installed in a metal case.

the surface of board. Solder the emitter lead of Q3 to the top of the PC board where it passes through ground foil. Install voltage regulator IC1 and check its lead orientation. The flat end should face the edge of the PC board.

Check all work done so far for accuracy and orientation of components. Trim all component leads close to the PC board. Solder all bottom foil connections. (Do not plug any unused holes with solder.) Carefully fabricate L2 (see Fig. 6) and install in the PC board. Use an 8-32 screw as a tool to hold L2 during installation. After installing L2, remove screw by unthreading it from the coil. Coat L2 with clear lacquer, Duco cement, or clear nail polish.

Use left-over lead clippings or bare, solid, tinned wire to install ground plane jumpers in

all open positions (see Fig. 4)—*this is very important*. Install a 1-inch length of bare tinned wire in +12v connecting hole near D8 to serve as a terminal. Place a piece of sleeving over it to avoid short circuits and make a small hook at the wire's end.

Install all surface-mount capacitors, as shown in Fig. 5, on the *bottom* foil of the PC board. Refer to detailed surface-mount instructions in sidebar *How to Install Surface-Mount Capacitors*. After installing all surface-mount capacitors, avoid flexing the PC board, which might crack the surface-mount capacitors.

Install the surface-mount resistors R14, 15, and 16 the same way as surface-mount capacitors (see Fig. 5). Resistors R14 and R16 are marked 221 (220 ohms) and R15 is marked 220

(22 ohms). Install the resistors with the marked side facing up from PC board so the markings are seen. Check the resistive values with an ohmmeter or DVM before installation.

Install Q1 as shown in Fig. 5. The thicker lead is the source (S) lead. Handle Q1 carefully; it is sensitive to electrostatic discharge (ESD). Check the orientation of Q1's leads before soldering in place. Then, install D1. The side of D1 with two leads is the ground side.

Tune-up and checkout procedures

The downconverter is aligned by peaking for maximum received signal. There is no tricky alignment procedure or specialized RF test gear necessary for good results and performance from the downconverter.

The successful tune-up of the downconverter requires either a known signal in the 900 or 1300 MHz band to act as a reference

and tune-up aid, or access to a frequency counter and a signal generator covering this range. A sweep generator is excellent if you can obtain the use of one. If you have an ATV transmitter, it can be used as a signal source—but remember to use a dummy antenna and keep the transmitter at some distance to prevent overloading the downconverter. You will need a suitable TV receiver or monitor tuned to either VHF channel 3 or 4, (an old B/W portable TV will do), a variable power supply supplying 10 to 20-volts DC at 40 milliamperes or more (preferably one with a built-in voltage and current metering), a DMM, and clip leads and cables as needed. If you cannot obtain these items, find someone to help you. A frequency counter that is reliable to at least 100 MHz higher than the downconverter receiving frequency is a great help in setting up the local oscillator and finding out where you are during tune-up.

First, carefully check your work for accuracy, component placement, and correct orientation, and poor solder connections. Make sure that surface-mount capacitors are correctly installed—a malfunction can destroy the gain of a stage. Also check for inadvertent solder bridges. When you are sure everything is OK, connect a DC source to the 12-volt input (D8 and ground). Ground is negative. No damage will occur should the power supply be accidentally reversed. Note the current drawn from the power supply. It should be around 35 ± 5 milliamperes. If higher than 40 milliamperes, check for possible short circuits. If lower than 30 mA, something may be open or missing.

Voltages checks

Refer to schematic diagram (Fig. 2) and parts layout diagrams (Figs. 4 and 5) so that you fully understand the circuit portions that will be tested. Unless specifically stated, the negative meter lead shall be grounded to negative terminal of power supply when making voltage measurements. If volt-

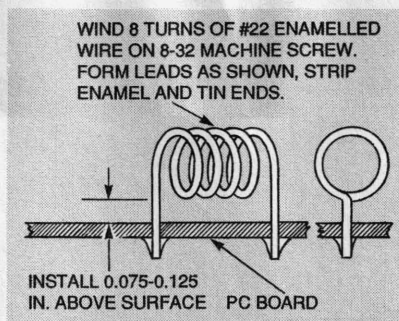


FIG. 6—COIL L2 FORMING and installation details. Wind 8 turns on 8-32 machine screw to form coil L2. Do not remove screw until after installation on PC board.

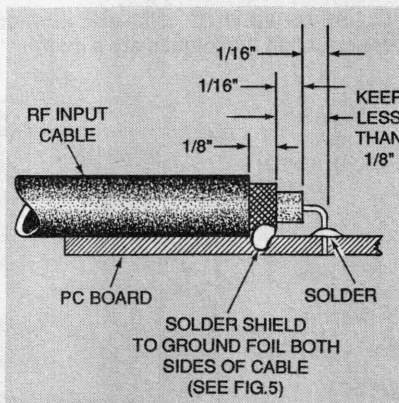


FIG. 7—INSTALLATION of RF input cable is critical. Keep recommended dimensions equal to or less than given in diagram, otherwise excessive losses will occur.

age discrepancies are noted, check the components indicated in Table 1 (page 69) for faults. Slight variations outside specified voltage limits specified may be ignored if component checks do not reveal anything wrong. If you experienced some deviation from the voltages measured but were not too far off, the downconverter may be OK, but remember and record these slight voltage deviations should any further problems be encountered. If trouble is experienced at a later step with that particular circuit, further investigation may be needed. Remember that the 8-volt on-board regulator has a 5% tolerance and this will affect all other readings. Also, your DMM could be a few percent in error as well.

1. Set the thumbwheel of R24 at its approximate center position, and set C22 so its plates are about $\frac{1}{4}$ meshed. Now measure the voltage at the base of

Q2. Using a non metallic tool, slowly rotate trimmer C22. There should be some perceptible voltage change at base of Q2 as C22 is rotated from minimum to maximum and back. Touching a metallic screwdriver blade to the stripline Z6 should also produce some change in this reading if Q5 is oscillating. A change of 0.01 volt or more is about what you will see if all is OK. This confirms that the local oscillator circuit is oscillating and that the mixer is getting a RF signal from the local oscillator.

2. Check that no part is getting hot.

3. If steps 1 and 2 are successful, connect the positive lead from a variable-voltage supply to junction C19 and L3, and connect the negative lead to ground. Set the wiper of R24 at ground (extreme counterclockwise rotation as viewed from shaft side). Set the power supply to +11 volts. Check for following voltages:

- At TP1 a voltmeter should indicate +7.6 to +8.4-volts DC as measured before in step 1. If not OK, check L3, D7, C27, and C28.

- At TP5 a voltmeter should indicate less than +1.2-volts DC. If not OK, check D6, D5, R27, R26, D4, and D3.

4. Increase the power supply voltage to 19 or 20-volts DC. Repeat step 3. The reading at TP1 should still read the same, however TP5 should be about +9-volts DC. If not, check all components mentioned in step 3. Steps 3 and 4 check out the remote tuning circuitry. If OK, proceed with the alignment in next step.

5. Connect the center conductor of a 75-ohm coaxial cable to the junction of C19 and L3 and the shield braid to ground as shown in Fig. 5. Terminate the end of the cable to J2, a connector suitable for your TV or monitor. (This is generally a type F connector.) Tune the TV to channel 3 or 4—whichever is unused in your area—and set its controls for normal reception. If you have a sensitive RF millivoltmeter, use it as an output indicator instead of a TV

set. The meter is easier to interpret.

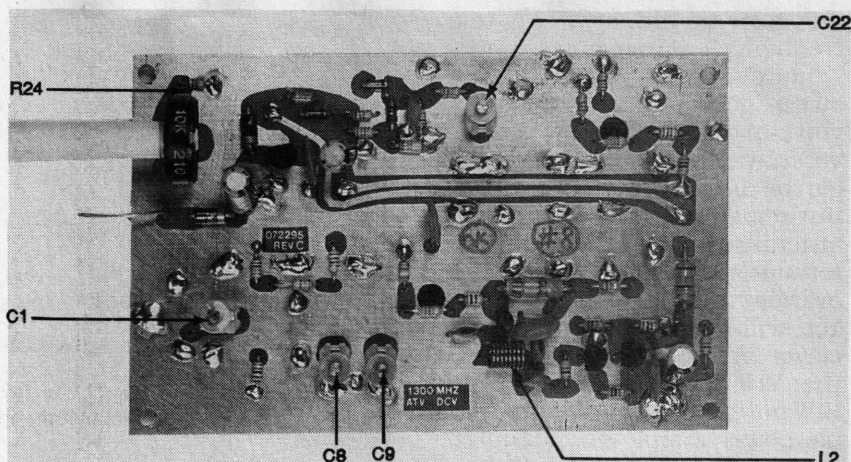
Connect a suitable signal source to J1, the downconverter input. Use 50-ohm low-loss cable and keep connections short. Refer to Figs. 5 and 7 for cable installation dimensions that are critical. The signal source can be either a generator or antenna. Preset all trimmer capacitors (there are four) to settings given in Table 2. Make sure to use the correct presets as the presets for 900 and 1300 MHz versions differ. Set the thumb-wheel of R24 to its center position.

6. Connect +12 to +14-volts DC from the power supply to D8, the negative lead connects to foil ground. Keep the bottom of the PC board at least 1/2 inch above any kind of material which may detune the microstrip elements on the bottom of the PC board. Because both conductors and insulators can cause detuning, a set of four 1/2 or 3/4-inch plastic or brass standoffs installed in the four corner holes of the PC board should be used. Use only a non-metallic tuning tool for all adjustments in the following steps.

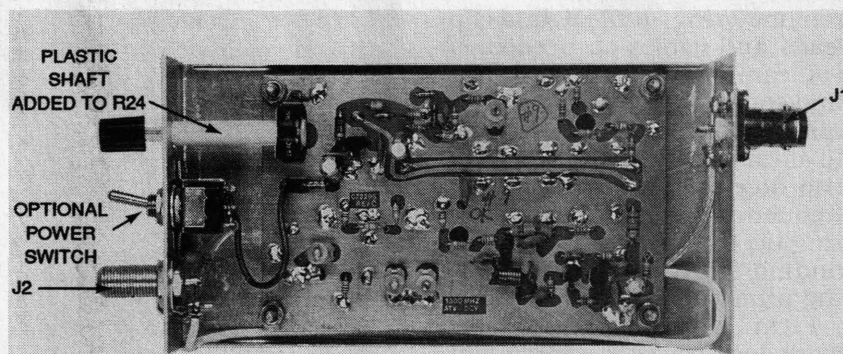
7. Activate the signal source and slowly rotate C22 until some indication of reception is seen on the monitor or RF voltmeter. Use a fairly strong signal at first. On confirmation of reception, decrease signal until it is just high enough for a reliable indication. As you remove the tuning tool from C22, undoubtedly some detuning will occur. You will have to compensate a little to get the tuning correct when the tool is removed. A frequency counter rated up to

TABLE 2
TRIMMER CAPACITOR PRESETS

Trimmer Capacitor	Percent of Mesh
Presets for 900-MHz model	
C1, C8, C9	25%
C22	80%
Presets for 1300-MHz model	
C1, C8, C9	10%
C22	20%



CLOSE-UP VIEW OF 1300-MHz downconverter board. Short lead lengths and neat construction techniques are a must!



MOUNT THE DOWNCONVERTER in a metal case for proper shielding. Use connectors to match your equipment.

1000 to 1500 MHz can indicate frequency shifts.

8. Next, couple the frequency counter to Q3. Stay at least 1/2 inch away from C22 and Q3. Set C22 for 850 ± 3 MHz with R24 at its center position (900 MHz model) or 1210 ± 3 MHz (1300 MHz model). You must have a steady reading, not a wildly jumping one. You may have to experiment with the coupling method, depending on your counter. We found loop coupling a little easier than the factory supplied whip antenna that came with our counter. Do not connect any probes, cables, or wires directly to the oscillator circuit or any of its components. A cable can be connected to the emitter of Q2 through a 220-ohm 1/8 or 1/4-watt resistor for the frequency counter, if desired, but this may detune the oscillator 10 or 20 MHz. However, with certain counters that are low in sensitivity above 800 MHz this may be the only way to get a steady reliable reading.

You can later compensate for this by touching up C22 once you are in the ballpark.

9. Once you obtain a steady reading and you are reasonably sure it is valid, proceed to check the local-oscillator tuning range. Rotate R24 through its entire range. For the 900 MHz model, the local oscillator should cover 840 to 867 MHz (assuming a channel 3 IF) and 1180 to 1240 MHz for the 1300 MHz model. Note that most ATV activity on 900 MHz is between 910 and 923 MHz, and in the case of the 1300 MHz, between 1240 and 1290 MHz (with gaps) so if you cannot quite achieve a 27 MHz range on 900 MHz, this should not be a problem. You can extend the range by replacing C23 on the 900-MHz model with a 5 pF chip NPO and re-tuning C22. The 1300 MHz model has somewhat wider tuning range (typically greater than 70 MHz). It is desirable to keep the tuning range as narrow as possible to improve the tuning

rate and ease of tuning a station. Adjust C22 as necessary to ensure desired coverage. For other IF frequencies, the local oscillator tuning range can be shifted lower (higher IF) or higher (lower IF). Note that the local oscillator must be *below* the signal frequency or else the IF signal will have its spectrum inverted with respect to the input signal spectrum. This will cause difficulties in TV reception in AMTV, and video and sync inversion in FMTV reception, unless the IF is designed for this.

If no frequency counter is available, you will have to depend on the reception of a known signal, or use a calibrated receiver covering the local oscillator range to pick up the local oscillator's signal.

If you don't have a frequency counter, receiver, or analyzer, don't worry. Use a known signal as follows. First set C1, C8, C9 and C22 to the presets shown in Table 2 if you have not yet done so. Set R24 at center of range. *Slowly* rotate C22 until some indication is seen on the TV or monitor, or other output indicator connected to junction of C19 and L3. Next, confirm that this is due to the signal source by shutting it off. If the indication disappears, this is the signal you want. Keep the signal level as low as possible for best results. If you get no results, make sure the signal has not gone off the air, and that trimmers are preset to correct settings, and the monitor, TV receiver, or indicating device is correctly set up and functioning. The presets are close enough to enable a strong signal to pass through the downconverter and give an indication.

10. Once you have an indication on the screen, peak C1, C8, and C9 for maximum signal as seen on the monitor. You should repeat this step as needed until no further improvement is obtained. Recheck setting of C22 to ensure correct tuning range is obtained. Check your final settings against presets. They should not be very different, as the tuning elements are printed on the PC board and therefore

tuning should be close to presets shown. However, if you are setting up the downconverter for other than the specified frequency ranges, the presets will vary.

11. Your downconverter is now working. Now you can go over the adjustments again to further improve performance. If you can get access to a sweep generator, align the downconverter for a flat response within ± 2 dB over the selected amateur band. The alignment is simple and straightforward so do not hesitate to experiment. You should be getting about 37 to 43 dB gain out of the downconverter and around 2 to 3 dB noise figure if everything is working properly.

12. The downconverter can be mounted in a shielded box with connectors of your choice. We recommend type N, TNC, SMA, or BNC for the input and an F connector for the output. Be sure to waterproof all connections for mast mountings. It is suggested that the alignment be checked after mounting the PC board in any enclosure since some slight detuning may result.

13. For remote tuning, make sure your variable supply is *clean*. Any noise or hum will cause FM on the local oscillator signal and a noisy received picture. Be sure also to set R24 fully CCW when remote tuning is used, or else the converter tuning range will be restricted or nil.

Variable gain

Both models of the downconverters develop 37 to 43-dB power gain from the antenna input to the RF output. This gain may be too much when strong TV signals are received, resulting in overload of the TV set. The unit's gain may be modified by changing R11 from the default value of 1000 ohms to other values as specified in Table 3.

A 1000-ohm potentiometer in place of R11 can be used for providing continuous variable gain. This portion of the circuit operates at the IF frequency, so a potentiometer and wiring tech-

TABLE 3—PRESET GAIN

Desired Gain (dB)	R11 Value (Ohms)
43	2,200
34	470
28	220
25	100

niques suitable at 70 MHz should be used. The IF output should be maintained below 150 millivolts rms in order to prevent overload of the converter. This is a very strong signal to feed into a TV receiver or video monitor.

Resistor R11 values assume a nominal gain of 40 dB. Increasing the gain beyond an extra 3 dB is not recommended. To reduce RF gain for very strong signals, it is recommended that a negative bias voltage supply of 0 to -10 volts be applied to TP7 through a 220,000-ohm resistor. This can be obtained from a potentiometer that is connected across a -10 V DC supply.

The potentiometer functions as an RF gain control in this case. Reverse AGC from the TV receiver used as a monitor could be arranged so as to bias TP7 from +1.5 volts DC at zero signal to -3.0 VOLTS DC on strong signals. Typically 35 to 40 dB RF gain reduction may be obtained using either of these approaches. As TV receivers vary, the individual circuit arrangements for doing this are left to the ingenuity of the builder.

Another gain reduction method is the use of a manually operated switch to ground TP7. This will reduce RF gain approximately 10 dB. This switch can be a simple SPST switch. For greater reduction of gain the switch may be connected to a negative voltage of -1 to -3 V as needed.

The matching ATV transmitter for the downconverter appeared in the May, 1996 issue of *Electronics Now* and corner reflector antenna construction will be covered in a future issue of *Electronics Now*. Refer to the May, 1996 issue for details of the system's overall performance in field trials. Ω

Modulating a Laser Beam

THIS MONTH, LET'S TURN OUR ATTENTION TO A DIFFERENT TYPE OF LASER APPLICATION—USING A LASER TO TRANSMIT SOUND. WE WILL LOOK AT A COUPLE OF DIFFERENT APPROACHES TO DOING THAT. EITHER APPROACH COULD MAKE

a great beginning for a winner of a "science-fair" project" or a starting point for further experiments.

Electronic Modulator

Our initial effort uses electronic modulation to send an audio signal over a laser beam. The technique we'll use provides a free-air transmission with a range of several hundred feet.

Of course, we need some way to decode the received beam and retrieve the signal. Take a look at the basic LM386 amplifier circuit shown in Fig. 1. To use that circuit as our laser "receiver," connect a silicon solar cell across the input and a small 8-ohm speaker or headset across the output. Now enclose the

circuit in a project box, and your receiver is ready. If you need to increase the range add a collimator tube at the input. It will also help improve audio clarity.

Now, how are we going to modulate the laser in the first place? The answer is another LM386 circuit. This time, we'll connect a microphone or another audio source to the input of the circuit in Fig. 1. To modulate the laser, we'll use the scheme illustrated in Fig. 2. As that circuit shows, the amp's output is connected to the 8-ohm winding of a small audio transformer. The 2000-ohm winding is connected between the laser's anode and the high-voltage power supply as shown.

With that set-up, the power fed to the laser will vary with the audio output of the LM386 circuit. Place the receiver so that the laser illuminates its solar-cell input, and whatever audio signal is introduced to the electronic modulator will be picked up and decoded by the receiver. Note that any audio-amplifier circuit could be used for this application,

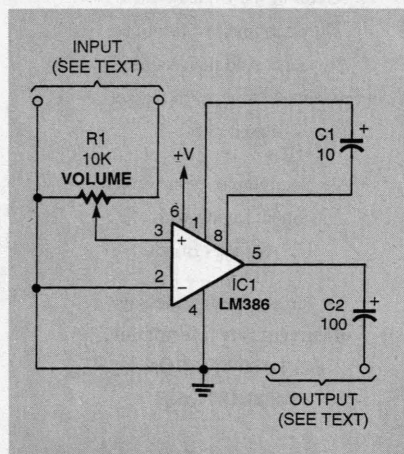


FIG. 1—THIS SIMPLE LM386 AMPLIFIER forms the basis for both a laser receiver and a laser modulator.

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but the LM386 circuit is simple, compact, and inexpensive—both units can be built for under five dollars.

Try using mirrors to deflect the beam around objects, or bend it around a corner. The laser beam can also be sent out to a reflector and back to the receiver. However you set it up, the system makes for an interesting demonstration.

Build the Photophone

Our second demonstration/experiment involves a mechanical modulation device often referred to as the Photophone. Invented by Alexander Graham Bell in 1880, he often expressed the opin-

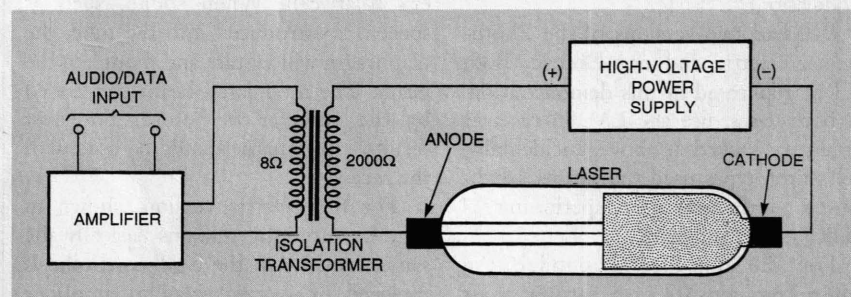


FIG. 2—USE THIS SCHEME when using the LM386 amplifier to modulate a laser.

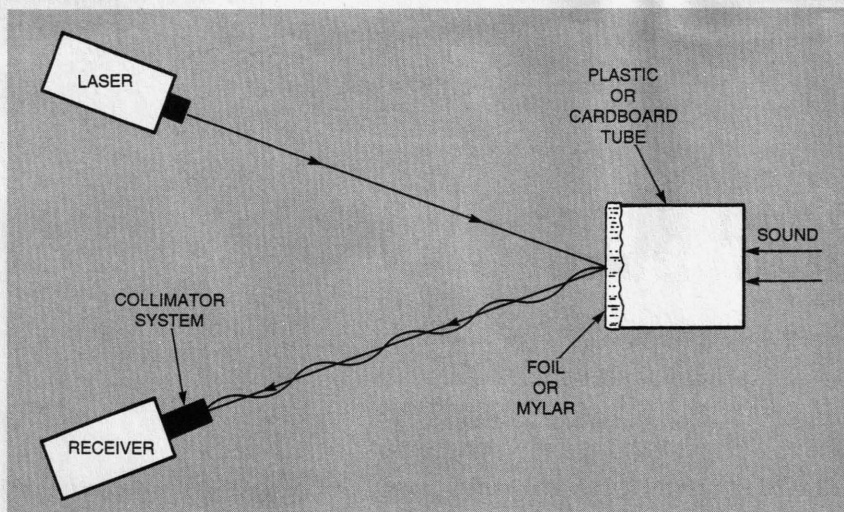


FIG. 3—BASED ON A PRINCIPLE developed by Alexander Graham Bell, this mechanical Photophone will modulate any beam of light, including a laser. The receiver is the same one used in our previous experiment.

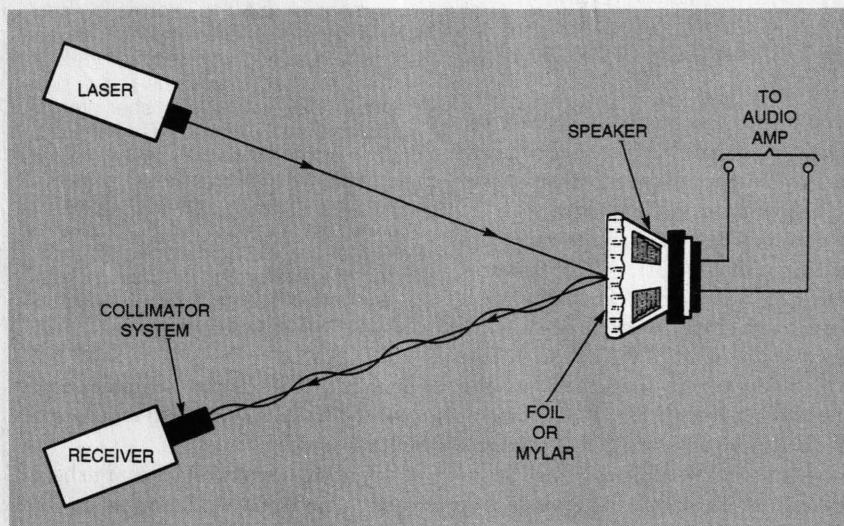


FIG. 4—SIMILAR TO THE MECHANICAL PHOTOPHONE, this electronic version replaces the tube with a speaker.

ion that the photophone was a more important invention than the telephone. Functioning on the same principle as the speaker modulator discussed in earlier installments of this column, the Photophone relies on the vibration of a metallic foil or Mylar diaphragm for the beam modulation.

Bell had two versions of the Photophone, electric and non-electric. Both will be replicated in this demonstration. In both cases, use the LM386 receiver circuit we looked at above. Incidentally, Bell's prototypes used the sun as a light source, but the laser is far superior in reliability, range, and clarity.

For the non-electric model, the transmitter consists of a cardboard or plastic tube with foil or Mylar stretched tightly over one end and secured with

tape or contact cement; see Fig. 3. The tube must be fastened to a tripod or other, similar support, so it will remain steady during use.

The laser and receiver are then set up so that the beam bounces off the center of the foil or Mylar and onto the receiver's solar cell. When sound, such as speech, is introduced into the tube, the diaphragm will vibrate and modulate the beam. The modulated beam is decoded by the receiver as before. For best results, a collimator should be used with the receiver.

For the electric version, shown in Fig. 4, everything remains basically the same, except that the cardboard tube is replaced by a speaker/audio amplifier assembly. The foil or Mylar is tightly stretched over the front of the speaker

and vibrates in sync with the speaker. Again, for best results, aim the laser at the center of the reflector and use a collimator with the receiver.

While innovative for its time, the Photophone is relatively simple in principle. Since its invention, numerous variations have been used, especially in the communications field. Can you think of any variations or experiments based on this principle? If you can, why not drop us a note and let us know. Write to Laser Editor, **Electronics Now**, 500 Bi-County Blvd., Farmingdale, NY, 11735 or via e-mail to lartronics@aol.com. We hope to hear from you soon!

EN

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Experiments With Laser Light

ONE OF THE MAJOR PURPOSES OF THIS MONTH'S COLUMN IS TO INTRODUCE FIRST-TIME LASER ADVENTURERS, AS WELL AS SEASONED EXPERIMENTERS TO SOME OF THE THOUSANDS OF WAYS THAT LASERS CAN AID, BENEFIT,

and even entertain you. Hands on experiments and demonstrations are great ways to display the wonders, and applications of this unique light form. We hope to give you a far better understanding of the laser and to offer many new ideas for its use. These should give you some ideas of your own and we invite you to share them with us. Send your ideas to *Laser Applications Editor, Electronics Now magazine, 500 Bicuty Boulevard, Farmingdale NY 11735, or via E-mail to lartronics@aol.com.*

Light beam modulation

The term Modulation is a controlled change of an otherwise constant signal to convey information or data. There are two ways to modulate a laser beam: mechanically and electronically. Let's explore the mechanical methods first, as they produce the most dramatic effects.

If you shine the beam onto a reflective surface that can be moved, the resulting beam traces patterns caused by that movement. One interesting way to do this is to get the surface to move in tandem with speech, music, or other sounds. There are commercially available X-Y modulators and axis generators that do a great job at

converting sound to patterns. They sell for \$100 and up. An easy and far less expensive approach is to use an old 3- or 4-inch, 8-ohm speaker, together with a 1½- to 2-inch length of flexible spring, and a ¾-inch square of thin but rigid aluminized mylar.

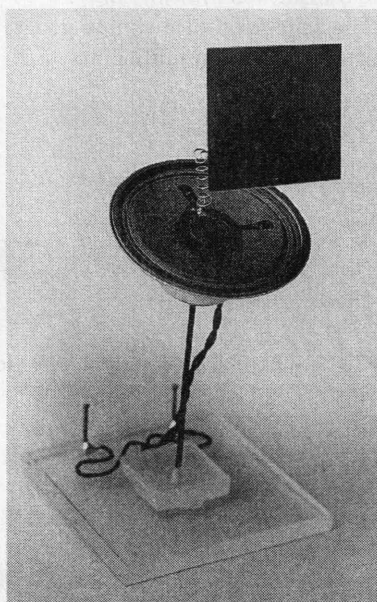


FIG. 1—A SPEAKER MODULATOR using flex-wire for positioning, and a small speaker with cone still intact.

You can use the speaker as is. I did, as you can see in Fig. 1. To reduce the sound output, which you

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might find disturbing, cut the cone out of the speaker, leaving the frame and the center circle covering the voice coil assembly. Use a dab of epoxy or silicon cement to attach one end of the spring to the center circle of the speaker—the less cement the better. When the cement is dry and the spring is in place, attach a corner or edge of the square of mylar to the loose end of the spring. Then mount the speaker on a stand with a swivel joint so the angle can be set anywhere within a 360-degree pattern, and you have a simple, but effective X-Y axis generator. A close-up photograph of the completed device is in Fig. 1.

To use your axis generator, connect the modified speaker to the output of a stereo, TV, radio, or other sound source. Then direct the angle of the speaker so that a laser beam aimed at the mylar creates a reflection that bounces back

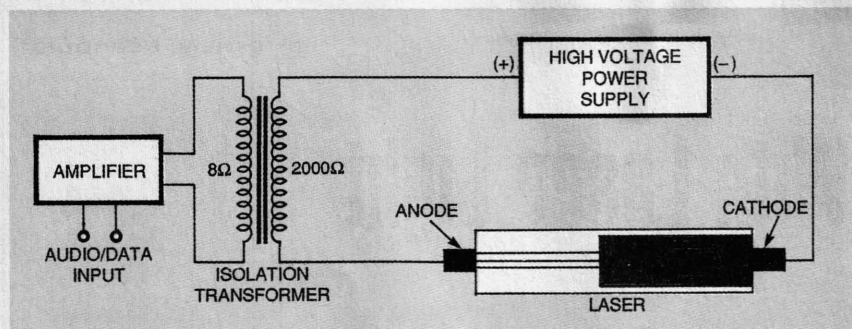


FIG. 2—ELECTRONIC MODULATION of a laser beam can be handled in this fashion. Obviously, the transformer you use must be able to handle the high voltage from the laser power supply without breaking down.

onto the surface you want it to strike—a movie screen, wall or ceiling. Now when music or some other audio signal is fed to the speaker, its center vibrates making the spring and mylar vibrate too. The reflected beam will swirl, circle, trace ovals, elongated shapes, and create broken and pulsating lines and patterns (no two designs ever seem to be quite the same). Viewing this produces hours of fasci-

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nation and relaxation. Even though you have removed the speaker cone, a bit of the audio will be heard. If you want greater volume, don't cut the cone out. If you want to muffle the sound more, put the speaker assembly into an insulated box and cut a small window for the reflected laser beam.

Electronic modulation is the second way to control the output beam. Figure 2 is a block diagram of the setup to vary the intensity of the laser light. The data, sound, or other information to be carried by the beam is fed into the amplifier's input. Its output is connected to the primary of the isolation transformer. The secondary winding is connected in series with the laser's anode lead, between the high-voltage power supply and the laser. The laser's cathode lead remains unchanged. With this arrangement the beam intensity stays constant

when there is no modulating signal. When a signal is applied to the amplifier, the impedance of the isolation transformer's primary winding changes and in turn, changes the impedance of the secondary. These changes vary the voltage applied to the laser and make the laser beam brighter or darker. In this way any form of information that can be converted into sound energy can be transmitted over the laser beam.

The amount of change is minute and often cannot be seen by the human eye. That's why a receiver consisting of a photosensitive transistor and amplifier is needed to decode the transmitted information. Figure 3 is the schematic of a simple general-purpose audio amplifier to use as

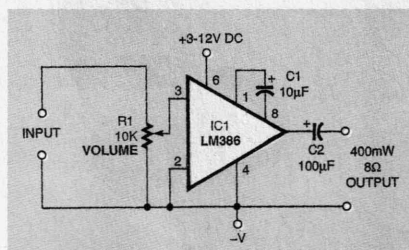


FIG. 3—SIMPLE IC AMPLIFIER for increasing gain of the received signal.

either the data input or data decoder. If you don't want to make your own amplifier you can buy a small modular style amplifier package at an electronics distributor. They are self-contained, complete with speaker and jacks for input and output.

Light-communications experiments

Once you have both a modulator and a receiver, you can use them to make an efficient "free air" light com-

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munications systems. Thanks to the intensity and collimation of the laser beam, these signals can be sent quite a distance, even in bright sunlight.

Beyond the line-of-sight free-air method of transmission, you can try a number of other interesting experiments involving reflecting and receiving laser light. For starters, try aiming the beam at various objects in and around your location. Buildings, windows, automobiles, signs, trees and other plants may amaze you with the variety of reflections they produce. Be sure and try this experiment both during the day and at night, as the results can be quite different. With the help of a friend, you can measure beam divergence or spread at various distances. Make a chart of your measurements and keep it with your laser.

Most 1-mW lasers will project a beam that will reach low clouds, up to about 2,000 feet. At night, the reflection and slight diffusion will create an eerie, almost UFO effect as the clouds move and change shape. Another fascinating observation calls for project-

Check Aug 1996 p.10

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GLOSSARY

anode—The positive (+) connection of a laser tube or diode.

ambient light—Existing light in and around a location.

beam—The narrow light emission from the anode of a laser.

beam splitter—An optical device used to divide the beam of light into two beams. Either prisms or partial mirrors.

cathode—The negative (-) connection of a laser tube or diode.

collimation—The property of laser light that keeps the beam from spreading out as it moves away from the laser.

collimator—A lens, or lens assembly that focuses the light into a beam.

coherence—The property of laser light where the beam emitted is largely of one frequency or color. In the case of the Helium-Neon laser, this color is red at 632.9 nm.

concave—The shape of a negative lens or mirror that allows for the spread of light. The surface curves inward.

convex—The shape of a positive lens or mirror that allows for the concentration of light. The surface curves outward.

diode—An electronic device that con-

ducts in one direction only. In some applications, it changes alternating current to a pulsating direct current. The positive side of the diode is the anode, and the negative side the cathode.

electrode—A metal contact point for electrical connection as in the anode and cathode of the laser.

electron—The negatively charged particles of an atom that orbit the nucleus. It's the action of the particles, jumping back and forth in the orbits, that creates the laser light.

fiber optic—Plastic or glass rods that transmit light from one end to the other. Allows for bending of the beam.

focus—Also referred to as focal point, it is the distance from a lens, or lens assembly, where the light comes to a point or apex.

helium—The elemental, inert gas with atomic number 2 and the symbol He.

hologram—A photograph, made with laser light that appears to have three dimensions.

holography—The science and technique of producing holograms.

infrared (IR)—A form of electromagnetic radiation, between visible red and the microwave range, with a frequency of 780 nm to 100,000 nm.

laser—An acronym for Light Amplification by Stimulated Emission of Radiation. This electrical or mechanical device produces a very straight and narrow beam of light, usually visible, of one dominant color.

nanometer (nm)—A term used to describe the wavelength of light, it is equal to a 1×10^{-9} meter.

neon—The elemental, inert gas with atomic number 10 and the symbol Ne.

optics—Any device used to control or manipulate light. Usually made of glass or plastic, these include lenses, mirrors, prisms, and filters.

photon—A quantum of light emitted by any source.

population inversion—The condition when more atoms are in an upper energy level than in a lower one.

refraction—The bending of light at the boundary of two surfaces.

spectrum—The entire range of electromagnetic energy. When applied to light, and/or lasers, it includes the frequencies from short wave ultraviolet through infrared energy.

ultraviolet (uv)—Light, invisible to the eye, from purple up, beginning at about 400 nm.

visible light—Electromagnetic radiation, from about 400 nm to 880 nm,

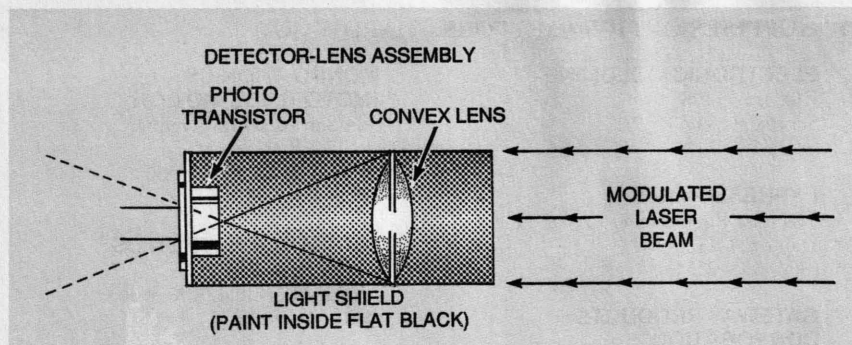


FIG. 4—DETECTOR-LENS ASSEMBLY is housed inside a tube that will fit the lens.

ing the beam through rain drops at night. Shining the beam out across a field during a mild rain shower is a very pretty sight, even during daylight. The same applies to fog. With the right angle, the beam will be visible only in the areas where it encounters the fog.

While you bounce the beam off various surfaces, you will probably notice that the light sometimes reflected back towards the source. The intensity of the reflection depends on the surface of the reflector. This property is what makes the next three experiments work. In each one a form of modulation comes into play. Once you can receive a returning beam of varying strength you can transfer information and, at the same time, demonstrate the principle of data transmission via reflector movement.

Because the reflected light is less intense, you will need a more sensitive receiver. You will also need an with considerably higher gain. The detector assembly has to be far more efficient in collecting the returning light and concentrating it on the phototransistor. The problem of light collection/concentration can be solved by using a combination light shield and

lens holder (see Fig. 4).

You can make one from either a cardboard, plastic, or metal tube. When properly focused, this lens condenses the collected light into a small spot on the photo-sensitive surface of the detector. If you need more sensitivity, use a cluster of phototransistors wired in parallel. Figure 5 shows how that is done. With the lens focused appropriately, this can be used to increase the photosensitive area. Paint the inside of the tube flat black to reduce ambient light interference. If this still does not adequately increase sensitivity, install a medium-

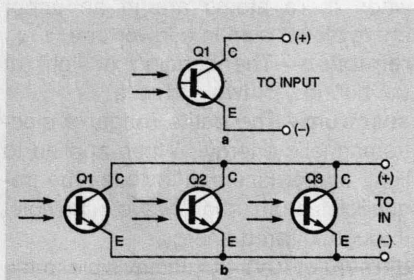


FIG. 5—SINGLE PHOTOTRANSISTOR used as laser modulation detector is shown in a. When more sensitivity is needed, multiple phototransistors can be connected as shown in b.

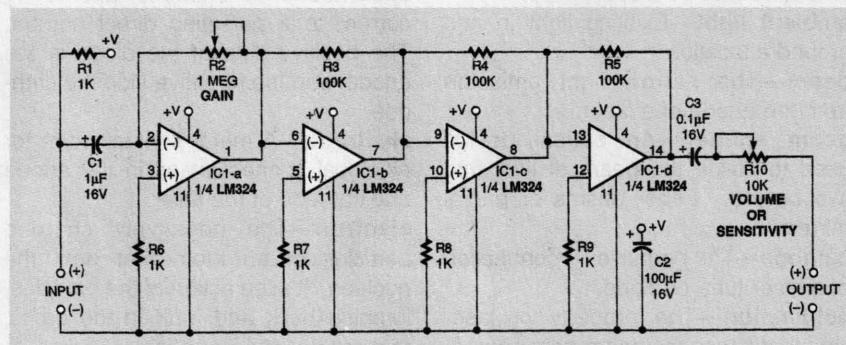


FIG. 6—FOUR-STAGE HIGH-GAIN amplifier uses a single LM324 quad amplifier.

density red filter at the front of the tube. Since the HeNe (Neon Helium) beam is red, it passes through the filter while green and blue light are rejected. This should reduce the ambient light to one third of its strength without the filter.

The amplifier shown in Fig. 6 is a four-stage, operational amplifier (op-amp) circuit wired in a cascade arrangement. Potentiometer R2 acts as the gain control for the initial amplifier stage. It serves two functions. First it helps set volume. Second it inhibits distortion that might otherwise be caused by exces-

sive gain. By feeding the output of the first stage into the inverting (negative) input of stage two, amplification is increased further. Stages three and four also boost gain. Potentiometer R10 is the final volume control for the signal being fed to the speakers or

headphones. The final output can also be connected to a relay or some alarm device. If this is done, R10 acts as a sensitivity control.

Once the detector and amplifier are completed, mount the LM324 circuit in a small project box. Position the detector assembly on the outside so it can be aimed at the returning light beam. The amplifier is powered by a standard 9-volt battery. Mount an on-off toggle switch and an "on" indicator, together with an output jack, on the project box.

You are now ready to start working on some practical projects.

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BUILD The GUARDIAN

You have invested a lot of your hard-earned money into your auto, van, or truck. Understandably, you'd like to protect it as much as you can while it is parked out of sight near your home or where you work. Some type of alarm system would certainly help in that respect. Most alarms with a siren don't bother thieves—they are so prevalent nowadays that a screeching alarm is ignored by almost everyone.

But if a vehicle is equipped with a silent alarm, thieves will take their time, giving the owner time to contact the local police. The Guardian alarm project presented here is just such a system, using state-of-the-art wireless technology. It is especially valuable to those who have recently purchased a new vehicle.

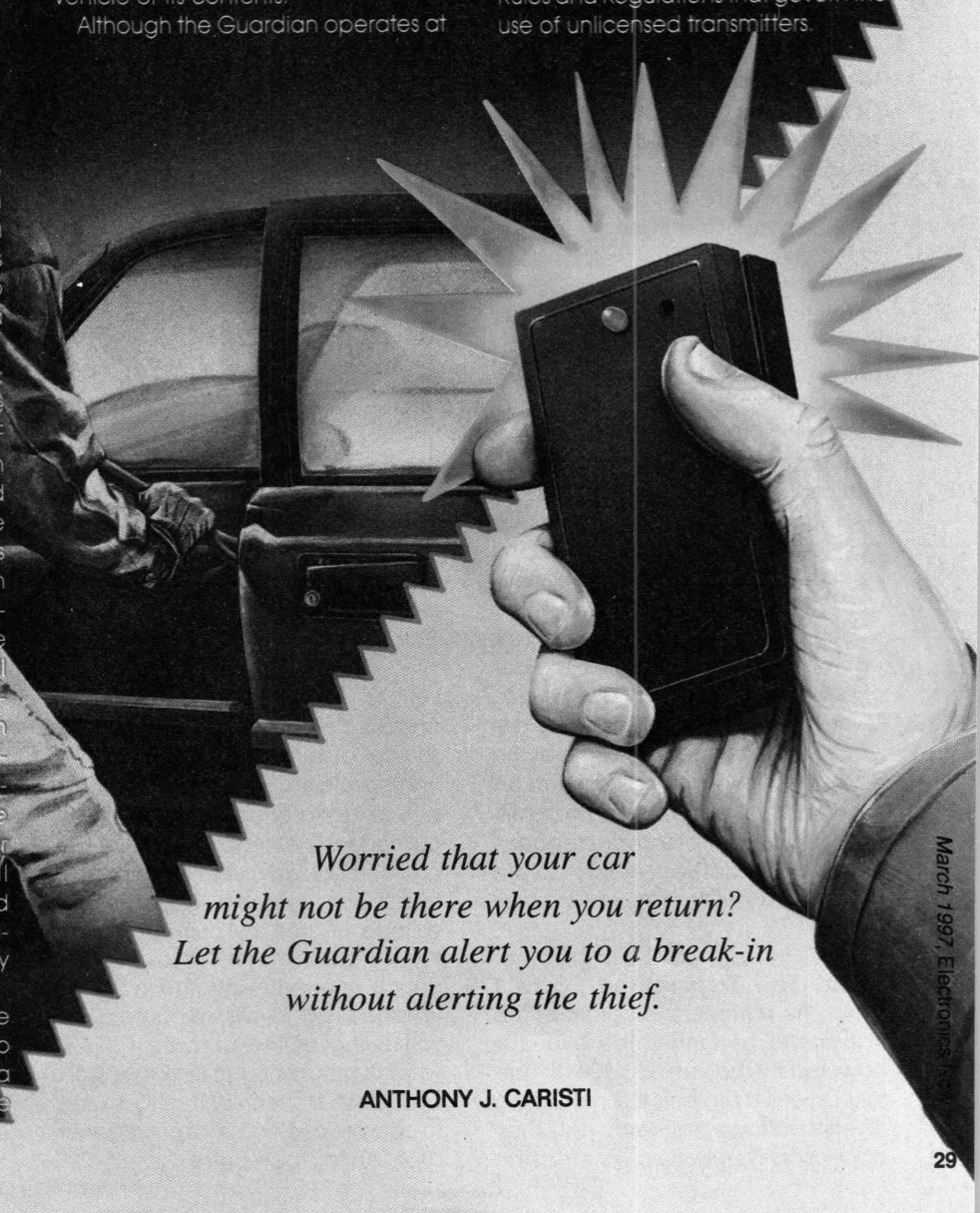
The Guardian consists of two parts: a transmitter and a receiver. The transmitter is a digitally-encoded UHF unit that is completely automatic in operation and permanently installed in the vehicle. Connected only to the 12-volt battery line, its current draw is just 2 milliamperes, so it can never run down the battery. The receiver, battery-operated for portability, might be carried around with you like a small pager, or placed at a convenient location and powered by a common wall-mounted transformer. The RF section of the receiver consists of a highly-sensitive, one-chip superheterodyne circuit that has been designed for easy construction using a small printed-circuit board. A second board contains the decoder and audio section, and the entire assembly fits neatly into a small enclosure.

Once installed in a vehicle, the Guardian is on duty at all times with no attention to it ever required. When a door on the vehicle is opened, the

Guardian senses the current flow to the vehicle's interior light, and automatically transmits a digitally-encoded UHF pulse train. That causes the receiver to respond with an attention-getting audible signal, so that corrective action may be taken before a would-be thief can steal the vehicle or its contents.

Although the Guardian operates at

433.92 MHz, it is easily built and no complicated RF-alignment procedures are required. The entire RF circuitry of the transmitter is contained on a hybrid module that is smaller than a postage stamp. That module has been designed to meet all requirements of Part 15 of the FCC Rules and Regulations that govern the use of unlicensed transmitters.

An illustration showing a hand holding a small, rectangular electronic device, which is the Guardian receiver. The device has a small circular speaker or light on its front face. In the background, a car is shown with its driver-side door open. A jagged, starburst-like graphic emanates from the receiver, suggesting it is active or transmitting. The overall style is a detailed black and white line drawing.

*Worried that your car
might not be there when you return?
Let the Guardian alert you to a break-in
without alerting the thief.*

ANTHONY J. CARISTI

March 1997, Electronics Now

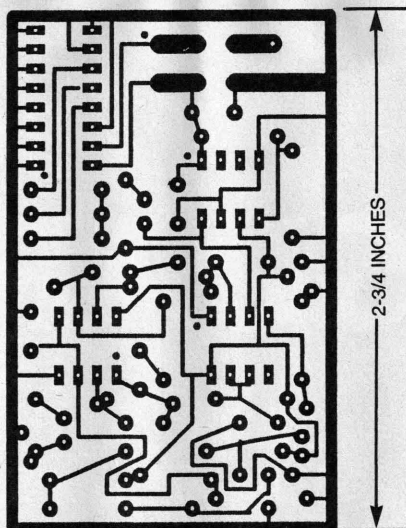
Digital Encoding And Decoding System.

A wireless-security system is useful only when the radio-frequency carrier is modulated by an encoded signal that is decoded at the receiver. That avoids spurious operation or false alarms due to interference or an unauthorized RF signal. The Guardian uses a sophisticated digital encoding/decoding scheme by using a pair of chips developed by Motorola. The transmitter contains the encoder, and the receiver contains the decoder. The two chips are specifically designed for wireless-signaling applications.

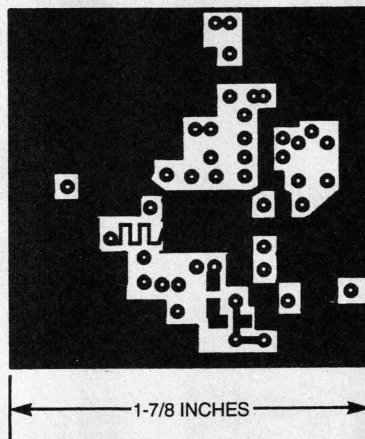
The encoder chip, an MC145026P, encodes nine digital bits of information that are programmed by means of hard-wiring to the input terminals of the chip. Each bit can be individually set to one of three states—logic 0, logic 1, or open. That permits up to 19,683 discrete codes. Both transmitter and receiver PC boards for the Guardian are hard wired for address zero, but any other address may be selected by re-connecting the addressing pins to any combination you choose. When the transmitter is enabled, the encoder chip generates a positive going pulse train at pin 15 that contains the encoded address. That pulse train is used to modulate the UHF oscillator.

Serial data is detected by the receiver and presented to the decoder at its data input terminal, pin 9. When that occurs, the address is checked. If two successive transmitted pulse trains contain the correct address, the valid output terminal of the decoder, pin 11, goes to logic 1 condition and remains so as long as the transmitter is operating. At all other times, the decoder output remains at a logic-zero level. The nine address bits of the decoder chip are wired the same as the encoder, allowing the receiver to respond only to its companion transmitter. If two or more independent systems are desired, each system should be wired for different address codes.

About The Transmitter. Figure 1 shows the schematic diagram of the transmitter assembly. The circuit is powered by the vehicle's 12-volt battery. Diode D1 protects the circuit from reverse voltage transients that may appear on the electrical system of the



Here is the foil pattern for the Guardian's transmitter. Clever layout allows for a single-sided board to contain both through-hole and surface-mount components.



The component side of the receiver board is mostly ground plane. The squarewave-like trace in the middle of the board is actually L6—necessary for proper operation.

vehicle. A pair of Zener diodes, D2 and D3, create two regulated voltage sources of 10.9 and 6.2 volts.

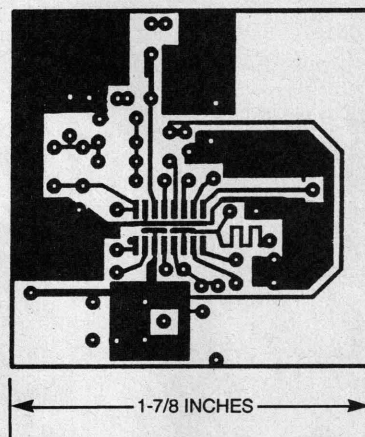
Integrated circuit IC2 contains a pair of identical op-amps, which are cascaded. The negative input of IC2-a is AC-coupled to the 12-volt bus so that it can detect a sudden sag in voltage caused by current draw of the vehicle's dome lamps when a door is opened. A change in battery voltage of 3 millivolts will cause a -6-volt swing at the output of IC2-b, which is great enough to trigger IC3, a CMOS 555 timer chip. That IC is wired to operate as a one-shot pulse generator. An output pulse about 2-1/2

seconds long appears on pin 3 of IC3 when triggered by IC2-b.

The output of IC3 is connected to the enable input of IC1, a switching-type voltage regulator. When IC1 is enabled, it outputs a regulated 3.3-volt DC power source for the transmitter portion of the circuit. That portion consists of encoder IC4 and transmitter module MOD1. When power is applied to the encoder, a series of pulse trains is generated that contain the address of the encoder. That pulse train is applied to pin 1 of the hybrid module to produce an RF signal with an on-off pulse modulation, sometimes called *amplitude-shift keying*. When the 2-1/2 second pulse time of IC3 is completed, the transmitter shuts down, returning to its dormant state.

In order to prevent RF transmission while the vehicle is running, voltage detector IC5 uses R11 and R12 as a voltage divider to sense when the supply voltage rises above 12 volts due to alternator charging when the engine is running. Under that condition, the output of IC5 is open, turning off Q1. That holds the reset input of IC3 low, preventing the timer chip from responding to any trigger pulses from the amplifier. As a result, no transmission takes place.

The HX1000 hybrid-transmitter module, MOD1, is a four-terminal device that contains a surface-acoustic-wave-stabilized UHF oscillator. The oscillation frequency, 433.92 MHz, is stabilized and controlled by the resonant frequency of the internal surface-acoustic-wave filters, which also filter out undesirable harmonics. An ad-



Here's the foil pattern for the solder side of the receiver board. Don't forget to solder connections on both sides of the board where necessary.

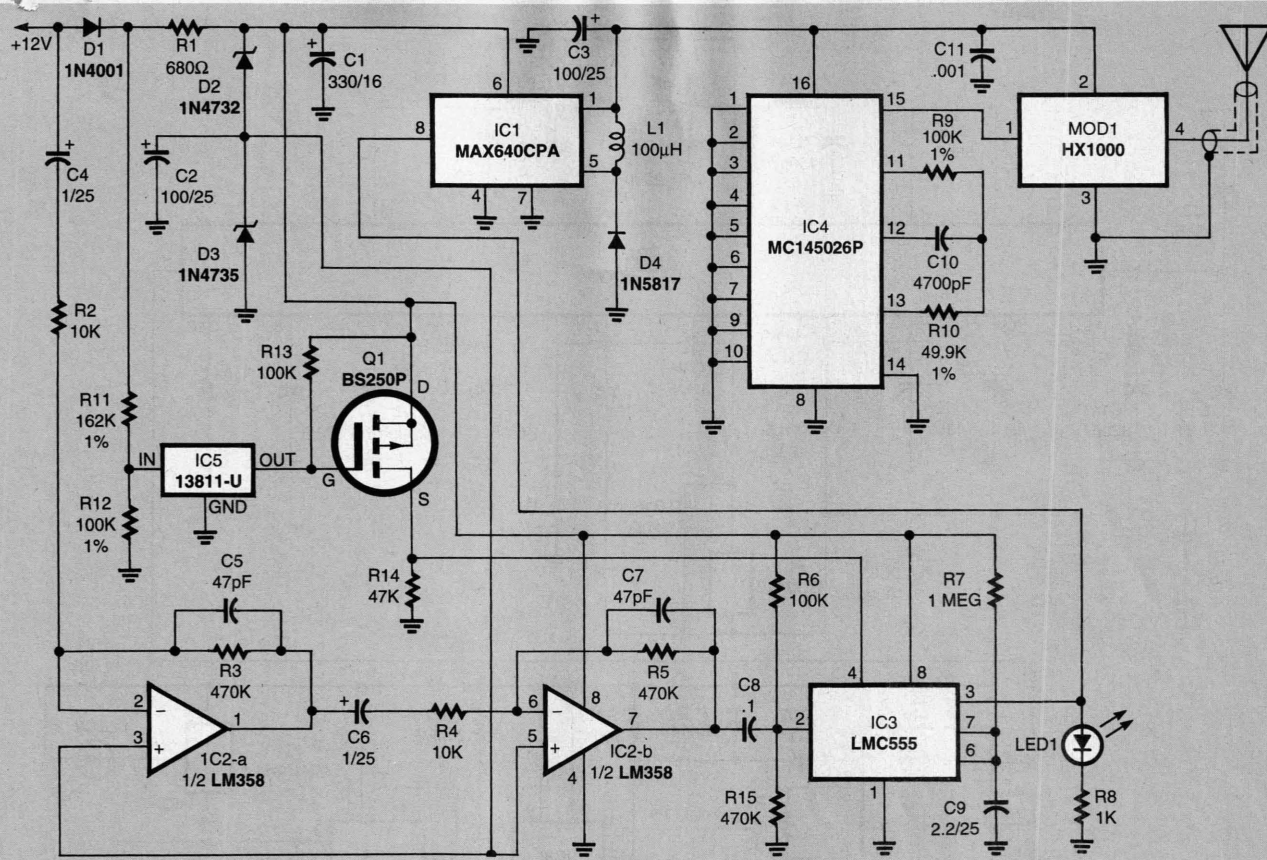


Fig. 1. Here is the schematic for the Guardian's transmitter. When a sag in the battery voltage is detected, an encoded address is broadcast to the receiver. Precision components and special ICs make the circuit adjustment-free and reliable.

PARTS LIST FOR THE GUARDIAN TRANSMITTER

SEMICONDUCTORS

D1—1N4004, silicon diode
 D2—1N4732A, Zener diode, 4.7 volts
 D3—1N4735A, Zener diode, 6.2 volts
 D4—1N5817, Schottky diode
 IC1—MAX640CPA, switching regulator, integrated circuit (Maxim)
 IC2—LM358N, dual op-amp, integrated circuit
 IC3—LMC555CN, CMOS timer, integrated circuit
 IC4—MC145026P, digital encoder, integrated circuit (Motorola)
 IC5—MN13811-U, voltage detector, integrated circuit (Panasonic)
 LED1—Light-emitting diode, red
 MOD1—HX1000, hybrid transmitter, 433.92-MHz (RF Monolithics)
 Q1—BS250PJ, P-channel, MOSFET transistor

RESISTORS

(All resistors are 1/4-watt, 5%, carbon units, unless otherwise noted.)
 R1—680-ohm
 R2, R4—10,000-ohm
 R3, R5, R15—470,000-ohm

R6, R13—100,000-ohm
 R7—1-megohm
 R8—1,000-ohm
 R9, R12—100,000-ohm, 1%, metal-film
 R10—49,900-ohm, 1%, metal-film
 R11—162,000, 1%, metal-film
 R14—47,000-ohm

CAPACITORS

C1—330- μ F, 16-WVDC, radial, aluminum electrolytic
 C2, C3—100- μ F, 25-WVDC, radial, aluminum electrolytic
 C4, C6—1- μ F, 25-WVDC, radial, aluminum electrolytic
 C5, C7—47-pF, ceramic-disc
 C8—0.1- μ F, ceramic-disc
 C9—2.2- μ F, 25-WVDC, radial, aluminum electrolytic
 C10—4700-pF, Mylar
 C11—0.001- μ F, ceramic-disc

ADDITIONAL PARTS AND MATERIALS

L1—100- μ H, inductor (Mouser 542-78F101 or similar)
 Hookup wire, 50-ohm coaxial cable (RG174 or similar), enclosure, PC board, etc.

vantage to using that module is that frequency tuning has already been done when the module was fabricated, so no further tuning is necessary. The hybrid module is capable of delivering about 1 milliwatt of power (0 dBm) into a 50-ohm load. It is connected to the transmitting antenna through a 50-ohm coaxial cable.

About The Receiver. The schematic diagram for the receiver is detailed in Fig. 2. Signals from the antenna are coupled through C12 to pin 14 of IC6, a complete superheterodyne-receiver chip that contains a local oscillator, mixer, limiter, and detector. A parallel-tuned circuit (L2 and C13) is adjusted to the carrier frequency of 433.92 MHz. The local oscillator is set to 433.42 MHz, and its frequency is stabilized by FL1, a surface-acoustic-wave resonator. The intermediate frequency (IF), which is the difference between the received frequency and the local oscillator frequency, is 500 kHz. The IF filter is composed of L3, L4, and C16 through C20. After IC6 processes the received RF signal, the

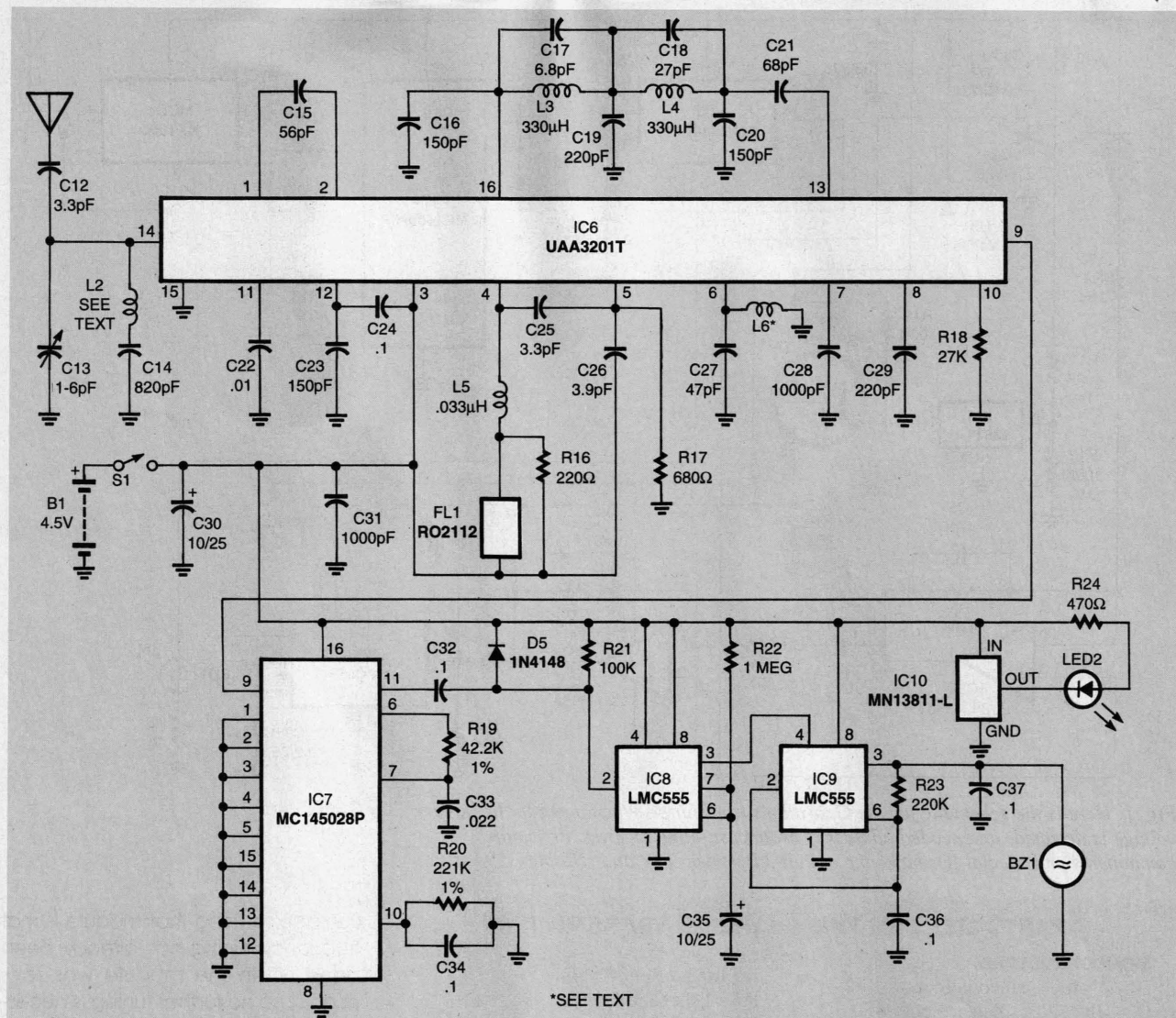


Fig. 2. The companion receiver for the Guardian takes advantage of the same reliable design as the transmitter. Only one simple adjustment is needed to tune the antenna to the Guardian's RF frequency.

transmitted pulses appear at pin 9. An unusual aspect of the design is L6, an inductor that is created by an etched pattern on the PC board. Because of that, use of the supplied PC pattern is an absolute must.

The companion decoder, IC7, checks the incoming pulse train for the correct address. If present, its valid output terminal, pin 11, goes high for the 2-1/2 second duration of the transmitted pulse trains. When the valid output signal returns to a low state, timer IC8 is triggered into operation. The output terminal of IC8, pin 3, goes high for a time period of about 11 seconds as determined by R22 and C35. That output, in turn, enables IC9, a timer configured as a 50%-duty-cycle oscillator with a period of about 1/4

second. The output of IC9 drives a piezo buzzer, which produces an attention-getting audio signal that alerts the user that a door of the vehicle has been opened.

Power to operate the receiver is obtained by a set of three "AAA" alkaline cells connected in series to produce 4.5 volts. Current draw is only 3 milliamperes during standby, so battery life will be in the hundreds of hours. IC10 is a voltage-detector IC that allows current to flow through LED2 when the battery voltage falls below 3 volts. That alerts the user that battery replacement is necessary.

Transmitter Construction. The transmitter is built on a small, single-sided PC board measuring about 1 3/4

by 2 3/4 inch. It may be placed into a small enclosure, and it is powered by the vehicle's 12-volt DC electrical system. A suitable length of 50-ohm coaxial cable is used to make the connection between the transmitter board and the Guardian's antenna.

Figure 3 illustrates the parts placement for the transmitter. For all components other than the RF module MOD1, the transmitter is assembled using standard through-hole wiring techniques. A printed circuit board is available from the source given in the parts list if you do not wish to etch and drill your own. Regardless of which source you use for the PC board, always clean the soldering surface with steel wool and detergent before assembly to remove any possible oxidation or dirt. That procedure will provide a clean metal surface and allow good solder joints to be made. Rinse the board in cold water and dry

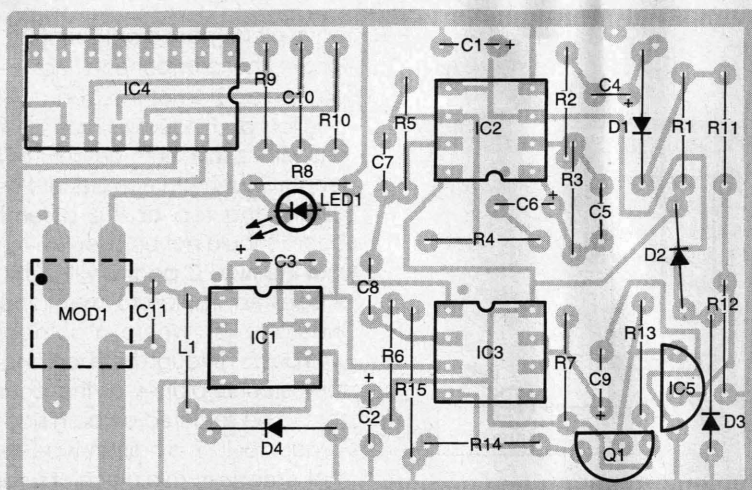


Fig. 3. Here is the parts-placement diagram for the Guardian's transmitter. A single-sided PC board makes the Guardian a surprisingly easy project to build. Only MOD1 is surface-mounted to the solder side of the board.

thoroughly before proceeding.

Insert all through-hole components in their respective locations as shown and solder them in place. It is important to use 1% tolerance metal-film

resistors as specified for R9-R12. Pay strict attention to the orientation of all solid-state devices and electrolytic capacitors. The ICs should be soldered directly into the circuit board

PARTS LIST FOR THE GUARDIAN RECEIVER

SEMICONDUCTORS

D5—1N4148, silicon diode
FL1—RO-2112, surface-acoustic-wave resonator, integrated circuit (RF Monolithics)
IC6—UAA3201T, UHF receiver, integrated circuit (Philips)
IC7—MC145028P, digital decoder, integrated circuit (Motorola)
IC8, IC9—LMC555CN, CMOS timer, integrated circuit
IC10—MN13811-L, voltage detector, integrated circuit (Panasonic)
LED2—Light-emitting diode, red

RESISTORS

(All resistors are 1/4-watt, 5%, carbon units, unless otherwise noted.)
R16—220-ohm, 1/8-watt, surface-mount
R17—680-ohm
R18—27,000-ohm
R19—42,200-ohm, 1%, metal-film
R20—221,000-ohm, 1%, metal-film
R21—100,000-ohm
R22—1-megohm
R23—220,000-ohm
R24—470-ohm

CAPACITORS

C12, C25—3.3-pF, ceramic-disc
C13—1.8–6.0-pF, trimmer (Mouser 24AA070)
C14—820-pF, ceramic-disc
C15—56-pF, ceramic-disc
C16, C20, C23—150-pF, ceramic-

disc
C17—6.8-pF, ceramic-disc
C18—27-pF, ceramic-disc
C19, C29—220-pF, ceramic-disc
C21—68-pF, ceramic-disc
C22—0.01-μF, ceramic-disc
C24, C32, C34, C36, C37—0.1-μF, ceramic-disc
C26—3.9-pF, ceramic-disc
C27—47-pF, ceramic-disc
C28, C31—1000-pF, ceramic-disc
C30, C35—10-μF, 25-WVDC, radial, aluminum electrolytic
C33—0.022-μF, Mylar

INDUCTORS

L2—Antenna coil (see text)
L3, L4—330-μH (Mouser 43LS334 or similar)
L5—0.033-μH, surface-mount (Mouser 434-07-033K or similar)
L6—See text

ADDITIONAL PARTS AND MATERIALS

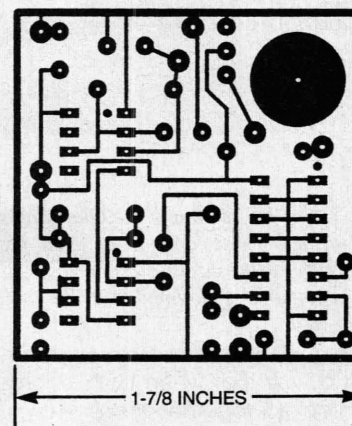
B1—4.5 volts, 3 alkaline cells, size "AAA"
BZ1—Piezo buzzer, 3–5-volt DC (see text)
S1—Slide switch, SPST, miniature (Mouser 102-1211 or similar)
Hookup wire, enclosure (Pac Tec K-HM-9VB or similar), PC boards, optional 5-volt DC wall-adaptor transformer supply (Digi-Key T309 or similar), etc.

for utmost circuit reliability since the Guardian is a security device. Once again, double-check the orientation before soldering—any polarized component placed backwards in the circuit will cause an inoperative transmitter and possible damage to one or more of the parts.

The RF module, MOD1, should not be installed at this time. It will be done later during transmitter test. That will ensure that the module is not subject to any voltage higher than 3.3 volts in the event there is an error in the board assembly. In the meantime, set the transmitter aside while the receiver is assembled.

Receiver Construction. The receiver consists of two printed-circuit boards that are stacked on top of each other and housed in a small plastic enclosure that contains a battery compartment. The receiver PC board is double-sided, and the decoder board is single-sided. Etched and drilled boards are available from the source specified in the Parts List if you do not wish to make your own boards.

Assemble the decoder board first, following the same precautions mentioned for the transmitter. Carefully orient all solid state devices and integrated circuits properly, following the parts-placement diagram in Fig. 4. Be sure to use 1% tolerance metal-film resistors for R19 and R20.



The decoder board is also single-sided. If you need to change the Guardian's address code, you can modify the traces by cutting them and attaching jumpers, or by changing the traces when etching the board. Don't forget to modify the transmitter board to send the same address!

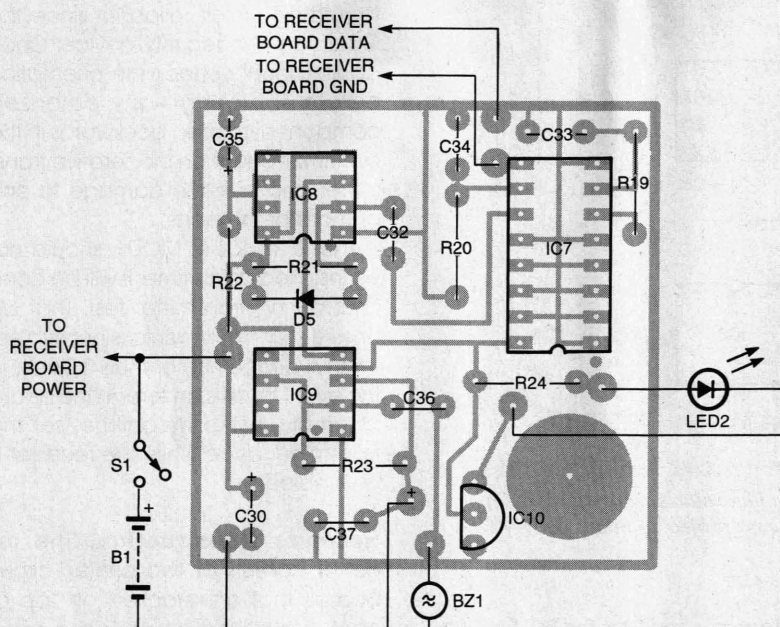


Fig. 4. The decoder section of the Guardian's receiver has an area reserved for mounting the buzzer directly to the board.

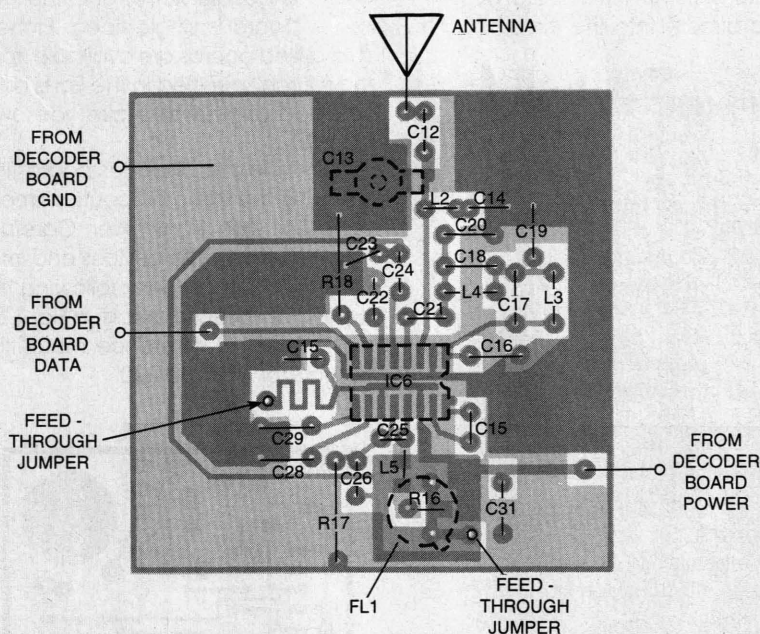


Fig. 5. The RF section of the Guardian's receiver uses a combination of through-hole and surface-mount components. Note the squiggly trace—it is actually an inductor necessary for the circuit's proper operation! The bottom side of the receiver board has three components mounted to it. Be careful of the orientation of IC6.

The body of BZ1 must be mounted in such a way that when both boards are stacked together, the entire assembly will fit into the enclosure selected for the receiver. If you are using the items recommended in the Parts List, drill a 1/2-inch diameter hole in the decoder board where BZ1 is to be located. If not, then drill a suitably-sized hole for the body of the buzzer you will be using. Cement the buzzer in place

using silicon rubber or epoxy. Be sure to observe proper polarity when wiring the buzzer to the circuit. For maximum sound level, drill a small hole or a series of holes in the enclosure where BZ1 is located.

The double-sided receiver board must be assembled using good RF construction techniques, and with the proper through-hole components as specified in the parts list. Use the

parts-placement diagram of Fig. 5 to locate the components that mount on the top side of the board. All capacitors and inductors must be inserted with a minimum amount of lead length. Maximum component height above the top of the assembled board should not be over 1 1/32-inch so that the two PC boards will fit into the suggested enclosure mentioned in the Parts List. Where a component wire passes through both top and bottom ground planes of the board, it should be soldered on both sides.

Inductor L2 is a hairpin-type design that is made from a piece of scrap 1/4-watt resistor lead. Use the dimensions in Fig. 6 to make the inductor. Solder surface-mount components L5 and R16 to the top side of the board. Note that one lead of C25 must be soldered on both sides of the board to make a connection for L5.

On the bottom side of the board, IC6, C13, and FL1 are placed as shown in Fig 5. Because of its unusual package design, the method used to mark pin 1 of IC6 is depicted in Fig. 7. When soldering IC6 to the board, center it within the foil pattern on the board. Carefully solder just one corner pin, and check the position of the chip to be sure all terminals are lined up with the pattern. If it is necessary to readjust IC6, it is easier to do so if only one pin is soldered. Once IC6 is properly oriented, solder the remaining terminals, being very careful not to use excessive heat or short one terminal to the next with solder. If inadvertent solder bridges occur during placement of IC6, use desoldering braid to remove the excess solder. Capacitor C13 is also surface-mounted to the board. Be sure to position C13 so that the screwdriver adjustment terminal is at ground potential. Filter FL1 mounts through holes in the board and is soldered from the top side.

Finally, two through-hole wires connecting the traces from the top of the board to the bottom are required. One is placed near FL1, and the other at L6 (the inductor created by the etched pattern on the PC board). Solder carefully on both sides.

Receiver Power Supply. For non-portable applications, the receiver may be operated by a wall-mounted power supply. Use a 4.5- to 5-volt, well-filtered DC supply with a load-current

PARTS LIST FOR THE DETECTOR PROBE

D6—1N90, diode
C38—4.7-pF, ceramic-disc capacitor
R25—10,000-ohm, 1/4-watt, 5%, carbon resistor

Note: The following items are available from A. Caristi, 69 White Pond Road, Waldwick, NJ 07463: Set of three etched and drilled PC boards, \$29.75; IC1, \$6.75; IC4, \$6.75; IC6, \$11.75; IC7, \$6.75. Add \$5.00 for shipping and handling. NJ residents add appropriate sales tax. MOD1 and FL1 are available from RF Monolithics, 4347 Sigma Rd., Dallas, TX 75244, Tel: 214-233-2903. Receiver enclosure is available from Pac Tec, 8425 Executive Ave., Philadelphia, PA 19153, Tel: 215-365-8400.

rating of 25 milliamps. It is important that the voltage applied to the receiver must never exceed 5 volts, or IC6 might be damaged. The Parts List specifies one suitable 5-volt DC wall-transformer supply. If you want to use a power supply that has a rating greater than 5 volts, you can add a 5-volt regulator circuit as shown in Fig. 8 to reduce the output voltage to a safe level.

Transmitter Test. The transmitter must be checked first so that it can act as a signal source to test the receiver. The test consists of two parts: First the analog circuits are checked for proper operation, then the transmitter module is installed and its RF output verified.

The transmitter will require a 12.6-volt power source that contains extremely low or no ripple in the output. A battery, not exceeding 12.6 volts, is recommended. If you use a DC supply that has a millivolt or more of AC ripple in its DC output, the detection circuit will trigger, making it difficult to test the transmitter properly. A voltmeter, either digital or analog, is needed to check voltages, and a triggered oscilloscope will be handy to observe transmitted waveforms. Additionally, a simple diode detector assembly may be constructed to verify transmitter operation. More about that later.

Apply the 12.6-volt power supply to the transmitter board, observing

proper polarity. The absolute maximum voltage allowed during the test is 12.8 volts. The measured voltage between the cathode of D2 and ground should be about 10.9 volts, and about 6.2 volts from the cathode of D3 to ground. If you do not obtain the correct readings, do not proceed until the fault is located and corrected. Components that should be examined carefully for proper value and orientation include D1–D3, C1, and C2. Examine the board carefully for shorts, opens, or cold solder joints, which may appear as dull blobs of solder.

Create a sag in the 12-volt power source by momentarily connecting a low value resistor across the battery terminals. You could use a 100-ohm 1-watt resistor or a 12-volt lamp for that. If you are using a regulated power supply, you could rapidly disconnect and reconnect the 12-volt lead to the transmitter. The sag in voltage should trigger IC3, and LED1 should light up for about 2-1/2 seconds. If LED1 does not light, check its orientation. You can test the operation of IC3 by momentarily shorting pin 2 to ground. That should cause pin 3 to rise

to about 10 volts for about 2-1/2 seconds, lighting up LED1. Also check that pin 4 of IC3 is at the same voltage as pin 8. If pin 4 is at a lower voltage, IC3 will think the car engine is running, and will not trigger. If that seems to be the case, check voltage detector IC5 and Q1.

If LED1 can be triggered manually, but does not sense a voltage sag, the trigger-output level at pin 7 of IC2B can be checked with an oscilloscope. A sag of about 3 millivolts on the input voltage is necessary to cause a negative going trigger pulse that goes to zero volts. If the amplifier is not working, check all components associated with IC2. A new chip might be necessary.

With LED1 reliably illuminating each time the battery voltage sags, measure the voltage at pin 1 of IC1. Normal indication is 3.3 volts, which lasts only as long as LED1 lights. If IC1 does not put out the proper voltage, check the orientation of C3, D4, and IC1. Check L1 for any assembly damage, as well as any opens or shorts in the PC board. You might have to replace IC1.

Temporarily connect the transmitter board to the vehicle's electrical system, and verify that LED1 lights each time a door is opened. The engine must be off when making that test.

With the analog circuits working, MOD1 may now be installed on the transmitter board. That component is surface-mounted on the foil side of the board. Before installing MOD1, all power from the tests should be removed. The four pins of MOD1 should be lightly tinned, using a minimum of heat from the soldering-iron tip. Additionally, the four pads on the PC board should similarly be coated with a very thin layer of solder. Doing so will greatly simplify achieving a solid solder connection between MOD1 and the foil pattern of the board.

Carefully note the location of pin 1 of MOD1 and pin 1 of the PC layout, both of which are identified by small dots. Using Fig. 3 as a guide, position MOD1 over the four copper pads so

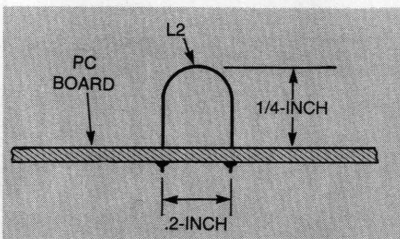


Fig. 6. In order to make L2, bend a scrap resistor lead to this size.

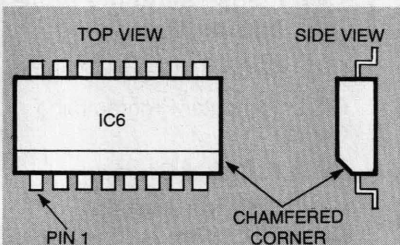


Fig. 7. The component used for IC6 has an unusual method of marking the location of pin 1.

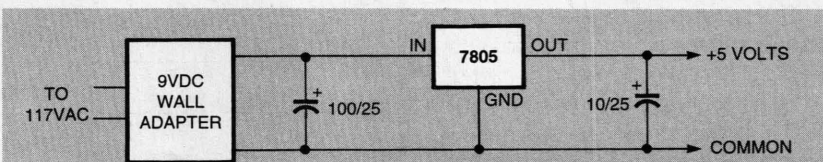


Fig. 8. If you want to mount the receiver at a permanent location, you can use this circuit to power the receiver from a wall outlet in place of the batteries

that pin 1 matches the corresponding PC pad, and the remaining three connections lie directly over their respective pads. Be careful with the soldering procedure to be sure the hybrid module is properly oriented and soldered in place. Use only enough heat and solder to attain good solder connections.

Detector Assembly And Final Checkout.

To test the RF output of the Guardian, a parallel-tuned detector probe is used to view the signal on an oscilloscope. Construction details for the detector are shown in Fig. 9. The tuned circuit of the detector probe is composed of C38 and an inductance that is composed of the capacitor leads formed into a 1/2-inch diameter circle and butt-soldered together. A germanium diode and resistor complete the circuit. To view the transmitted RF pulses, an oscilloscope probe set to a 1:1 attenuation is connected across the resistor as shown.

Cut a piece of insulated solid wire about 6-1/2 inches long and connect it to the antenna pad (pin 4 of MOD1) to act as a temporary antenna. Apply 12.6-volt DC power to the transmitter (as before, do not exceed 12.8 volts) and trigger IC3 either manually or by sagging the power supply. Note that LED1 lights for a couple of seconds.

Adjust the scope for maximum DC gain and place the pickup loop of the detector probe near the antenna. Set the sweep speed of the scope to 10 milliseconds per centimeter. When LED1 is lit, you should be able to see a display of 18 positive-going pulses, indicating that the transmitter is sending out a modulated RF signal.

If you do not see the pulse display, check the detector probe to be sure it is properly assembled. Check pin 15 of IC4 with the oscilloscope probe (set to 10:1 attenuation) to verify the presence of the 3.3-volt peak-to-peak pulse train. If IC4 is not generating a pulse train, check the orientation of IC4 and Q1, along with R9, R10, R11, R12, and C10. The encoder chip may have to be replaced. If the pulse train is normal, check the orientation and solder connections of MOD1.

Once all tests on the transmitter board are done, it may be installed in a small enclosure that has been drilled to accommodate two power leads and a coaxial cable. Use two

different colors for the power leads. Red for plus and black for ground makes the best choice, as those colors are almost universal for automotive electrical systems.

Receiver Checkout. Once the transmitter is working properly, it is used to test the receiver. The assistance of another person will be handy in testing the receiver at a dis-

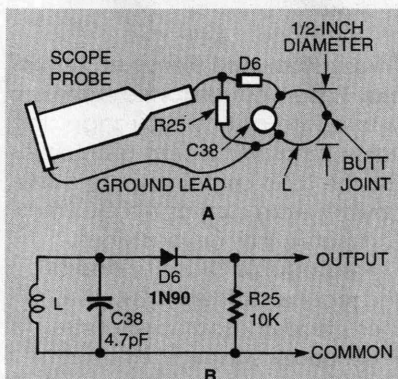


Fig. 9. This simple detector probe will allow you to view the Guardian's RF signal on an oscilloscope. Follow the dimensions and shape of the leads for C38 exactly as shown in A; the probe's schematic is shown in B.

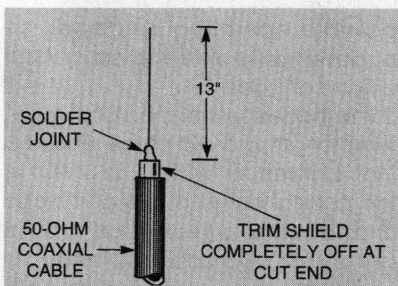


Fig. 10. A simple length of wire forms a 1/2-wavelength antenna for the transmitter.

tance from the transmitter. If you need to, the transmit time may temporarily be increased by connecting additional capacitance across C9 in the transmitter. Connect a 6-1/2 inch insulated solid wire antenna to the RF-input terminal of the receiver printed circuit board. Connect the probe of an oscilloscope to pin 9 of IC6. Apply 4.5 volts DC power to the receiver board, using either a set of three alkaline cells connected in series or a well-regulated DC power source. Do not power the receiver with any voltage greater than 5 volts, or you might damage IC6.

When the transmitter is triggered, the oscilloscope display should show

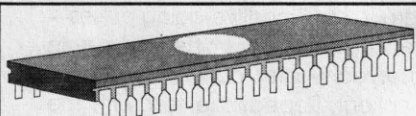
a group of 18 positive-going pulses—a replica of the transmitted pulse train observed earlier during transmitter checkout. Repeat the test with the transmitter further and further away. When the transmitter is far enough away to deliver a weak signal to the receiver, adjust C13 to obtain the maximum possible pulse width as indicated by the scope. The transmitter will have to be moved a substantial distance away to obtain a weak signal at the receiver. If you are not able to receive the transmitted signal, the receiver RF board must be visually checked for improper assembly. Every solder joint must be examined, especially those for IC6, L5, R16, and FL1. Verify that the jumpers between the top and bottom of the board near FL1 and L6 have been properly soldered. Check for short circuits and proper component values.

Local-oscillator operating current can be verified by measuring the voltage at pin 5 of IC6. Normal indication is about 0.5-volt DC. A reading of zero means that the local oscillator is not working. If that is the case, check L5, C25, C26, C31, R16, R17, and FL1.

When the RF board is operational, connect it to the decoder board. Three connections are needed: power, ground, and the data line. Disconnect the power supply to the receiver board and follow the parts-placement diagrams for the receiver and decoder boards using 24-gauge stranded insulated wire. If possible, use different colors. Do not use solid wire—it will break.

Disconnect any temporary capacitor connected across C9. Apply power to both the receiver and transmitter. It is OK if the buzzer sounds when the receiver is turned on. Move the transmitter across the room and trigger it as before. Each time LED1 goes out, the buzzer should emit a series of rapid beeps for about 11 seconds. The audio signaling time may be lengthened by increasing the values of either R22, C35, or both.

If the buzzer is silent, check pin 11 of IC7 to verify the existence of the valid data output signal, which remains at 4.5 volts when the transmitter is operating, and zero when it is not. If IC7 is not recognizing the pulse train, check the IC and all components associated with that chip. Verify that the address selected for both encoder and



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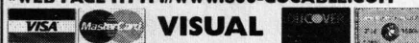
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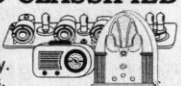
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decoder chips match. If the valid data signal is present, check IC8, IC9, and all components associated with those ICs. Manually trigger IC8 with a momentary short between pin 2 and pin 1 (ground). That should cause pin 3 to rise to about 4 volts for a period of 11 seconds. If both IC8 and IC9 work correctly, check the buzzer wiring and the buzzer itself. Once the receiver passes all tests, it may be assembled into the enclosure.

Transmitter And Receiver Antennas. Since they control the maximum attainable operating range, the transmitting and receiving antennas are the most critical parts of a low-power system such as the Guardian. Fortunately, they are easily built.

For portable receiver operation, the receiver antenna is simply a 6-1/2 inch piece of insulated solid wire that is connected to the RF input terminal of the receiver board, and bent to fit inside the periphery of the enclosure. The length chosen is a 1/4-wavelength antenna. For non-portable applications, a 1/4-wave or 1/2-wave (13-inches long) vertical-ground-plane antenna could be used. Note that the greatest possible operating range will be achieved using a well-constructed receiving antenna.

The transmitter antenna is a 1/4-or 1/2-wavelength ground-plane antenna that is driven by MOD1. The shield of the coaxial cable and the vehicle's chassis make a very good ground plane. Most applications will need the transmitter to be placed a short distance away from the antenna. A length of 50-ohm coaxial cable must be used to make the connection. Use Fig. 10 as a guide when constructing the transmitting antenna. Note that extremely short connections must be used when attaching the center wire and shield of the cable to the RF output and common pads of MOD1.

Transmitter Installation. The best place to locate the transmitter is near or under the dashboard of the vehicle so that the length of coaxial cable is short. The enclosure should be secured to the vehicle so that it does not move around when driving. Best operating results will probably be obtained with the transmitting antenna placed in a vertical direction at the center of the windshield of the vehicle. The wire

may be secured to the glass using transparent tape or even a quick-setting epoxy or instant-bond glue. Before mounting the antenna permanently, make sure it does not interfere with the driver's view of the road ahead—most, if not all, states have laws about just such a situation. If you wish, other locations for the transmitting antenna may be selected. A little experimentation with position and antenna length will indicate which location in your vehicle will provide the best range for the system.

Power to operate the transmitter is obtained from any wire in the electrical system powered by the 12-volt battery, so that the circuit is energized when the vehicle is parked. In most vehicles, the best selection will be the hot lead that feeds the dome lamps. The ground lead of the transmitter may be connected to any metal part of the vehicle.

Using The Guardian. The transmitter is always on duty, so no attention to it will ever be required. It will automatically transmit a burst of coded pulses each time a door is opened. It will even make one last transmission if a would-be thief disconnects the battery of the vehicle!

It would be a good idea to have an assistant open and close a door of the vehicle several times while operation of the receiver is checked at different locations. This will provide information as to the range of the system. Tests on the author's prototype demonstrated a useable range in excess of 200 feet for line of sight transmissions, using the portable receiver.

Battery-operated receivers should be turned on only when it is desired to monitor the status of the vehicle, such as when parked nearby out of sight. Upon power-up, the piezo buzzer may emit a series of beeps. When the receiver becomes silent, it is ready to receive a transmission. The receiver's LED indicator is a low-battery warning and it will not illuminate until it is time to replace the battery. If the receiver is powered by a wall-adaptor supply, it may be left on continuously so that the vehicle is protected at all times when it is parked nearby.

That's all there is to it! Why not start building your Guardian today so that it can begin watching over your car as soon as possible.

Ω



Q & A

READERS' QUESTIONS, EDITORS' ANSWERS
CONDUCTED BY MICHAEL A. COVINGTON, N4TMI

Long-Period Timer

Q Is there any chip or simple circuit that can be used as a reliable timer for a long time delay (30 days) with good repeatability? — J. Z., Mississauga, Ont., Canada

A Let's see, 30 days equals 2,592,000 seconds, which means you'll need to count billions of cycles if you use a crystal clock oscillator. The cheapest way is to use a microcontroller, which can count up to any number by using multiple registers. Each 8-bit register can count to 255; each time it rolls over, increment another register, and chain registers together until you can count high enough.

Off-the-shelf long-period-timer chips also exist. The 4060 includes a crystal oscillator and 14 divide-by-two stages so that it can count to 16,384; it runs on supply voltages from about 5 to 15 volts. The 74HC4060 is an equivalent device except that it requires a fixed 5-volt supply. Three of those chips in cascade, controlled by a 1.696-MHz crystal, would produce a square wave with one cycle every 30 days. Figure 1 shows a circuit using that approach. Unfortunately, we did not have a few months to spare, so we haven't tested it fully. You can get custom-made crystals from JAN Crystals, PO Box 60017, Fort Myers, FL 33906, and other suppliers.

frequency of oscillation. Note that overtone crystals will oscillate at their fundamental frequency; for example, 27-MHz CB crystals will oscillate at 9 MHz, and most scanner crystals will oscillate around 18 to 20 MHz.

caption built in, but in all the schematics I've looked at, the decoding is an integral part of the master processing chip. Is there a chip that can easily be added internally to older sets so that a decoder box doesn't have to be carried around and connected to different sets around the home? If so, where can I get information on the chips and associated circuitry? — J.N.B., Denver, CO

Closed-Caption Decoder

Q I've been told that the chips for decoding TV closed captions cost about \$5. I know all modern TVs are required to have closed-

A It isn't simple and certainly isn't a matter of a single off-the-shelf chip. You need a microprocessor that has access to

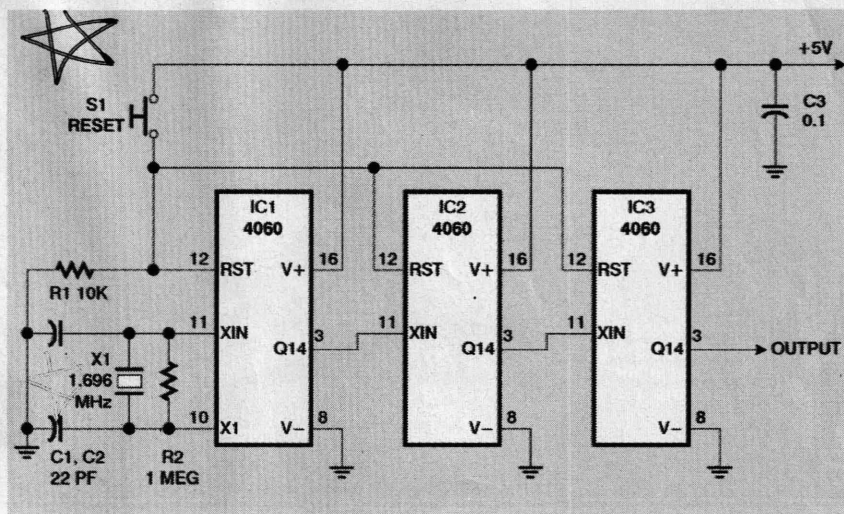


FIG. 1—THIS SUPER-SLOW OSCILLATOR produces a squarewave with a period of 30 days (one cycle per 30 days). The crystal can be ordered from the source provided in the text.

Crystal Tester

Q I am building a test panel and would like to include a tester to tell whether crystals are good or not. Can you help? — R. B., Langley, B.C., Canada

A Figure 2 shows a classic Colpitts oscillator circuit that will test crystals from about 1 MHz up. The LED glows (not at full brightness) when the crystal is oscillating; for best results, use a low-current LED. You can connect a frequency counter across the LED to measure the

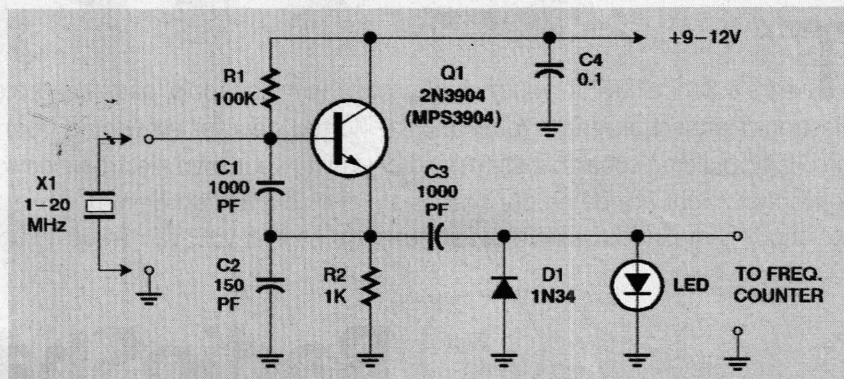


FIG. 2—THE VENERABLE COLPITTS OSCILLATOR can be used to test crystals from 1 to 20 MHz, and even higher. If the crystal is good (oscillates), the LED will light.

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the video and sync signals and the ability to generate its own video (for the lettering) in sync with what's coming in. The only reason closed-caption decoding is cheap to add to modern TV designs is that they already have such a microprocessor for other purposes, such as displaying the channel number and setup menu on the screen. So the closed-caption decoder isn't really a chip, it's some additional software in the chip that's already there.

See our December, 1994, issue, pp. 31-43, for a closed-caption decoder that sends its output to a personal computer rather than the TV screen. See also <http://www.broubaha.com/~eric/pic/caption.html> for a circuit and the code for closed-caption decoding with a PIC microcontroller.

Safe NiCd Discharge

Q I found out by accident that letting a NiCd battery pack sit unused for a long time (several months or more) safely "conditions" the power pack to accept a full charge as if it were new. In time, the cells discharge to a very low voltage without the danger of reverse-polarizing a cell, which is what might happen during deep discharging with a load. If I used a very light load (say 1 mA), would it be possible to discharge a battery pack safely without reverse-charging any of the cells? — J. A., via e-mail

A It's well known that if you charge a NiCd battery a lot, but never discharge it deeply, its performance will suffer. On the other hand, if you discharge the battery below about 1 volt per cell, you risk reverse-charging and ruining the cells that go dead first, since current will still be flowing through them from the other cells.

Leaving a battery pack alone is a good way to deep-discharge all the cells without reverse-charging any of them, but as you've discovered, it takes a long time. Unfortunately, we know of no shortcut. When you leave the battery unused, each cell discharges through its own internal leakage without sending any current through the other cells. If you use a load, no matter how light, the current passes through all of the cells regardless of each cell's state of charge. As you surmise, very light loads are better than heavy loads, and a 1 mA load might actually be quite safe—but full discharging would still take a couple of months.

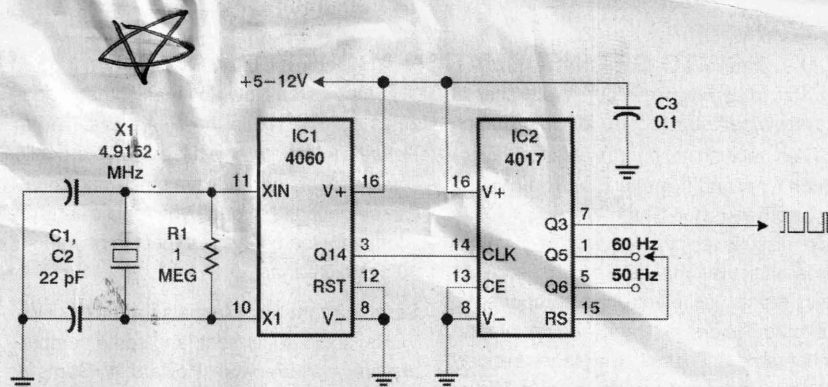


FIG. 3—THIS TWO CHIP CIRCUIT can be used to generate an accurate 50 Hz or 60 Hz clock signal using a standard microprocessor crystal.

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On the Internet: See our Web site at <http://www.gernsback.com> for information and files relating to our magazines (**Electronics Now** and **Popular Electronics**) and links to other useful sites.

To discuss electronics with your fellow enthusiasts, visit the newsgroups sci.electronics.repair, sci.electronics.components, sci.electronics.design, and rec.radio.amateur.homebrew. "For sale" messages are permitted only in rec.radio.swap and misc.industry.electronics.marketplace.

Many electronic component manufacturers have Web pages; see the directory at <http://www.hitex.com/chipdir/>, or try addresses such as <http://www.ti.com> and <http://www.motorola.com> (substituting any company's name or abbreviation as appropriate). Many IC data sheets can be viewed online. Extensive information about how to repair consumer electronic devices and computers can be found at www.repairfaq.org.

Books: Several good introductory electronics books are available at RadioShack, including one on building power supplies.

An excellent general electronics textbook is *The Art of Electronics*, by Paul Horowitz and Winfield Hill, available from the publisher (Cambridge University Press, 1-800-872-7423) or on special order through any bookstore. Its 1125 pages are full of information on how to build working circuits, with a minimum of mathematics.

Also indispensable is *The ARRL Handbook for Radio Amateurs*, comprising 1000 pages of theory, radio circuits, and ready-to-build projects, available from the American Radio Relay League, Newington, CT 06111, and from ham-radio equipment dealers.

Copies of past articles: Copies of past articles in **Electronics Now** and **Popular Electronics** (post 1993 only) are available from our Clagck, Inc., Reprint Department, P.O. Box 4099, Farmingdale, NY 11735; Tel: 516-293-3751.

Electronics Now and many other magazines are indexed in the *Reader's Guide to Periodical Literature*, available at your public library. Copies of articles in other magazines can be obtained through your public library's interlibrary loan service; expect to pay about 30 cents a page.

Service manuals: Manuals for radios, TVs, VCRs, audio equipment, and some computers are available from Howard W. Sams & Co., Indianapolis, IN 46214 (1-800-428-7267). The free Sams catalog also lists addresses of manufacturers and parts dealers. Even if an item isn't listed in the catalog, it pays to call Sams; they may have a schematic on file which they can copy for you.

Manuals for older test equipment and ham radio gear are available from Hi Manuals, PO Box 802, Council Bluffs, IA 51502, and Manuals Plus, PO Box 549, Tooele, UT 84074.

Replacement semiconductors: Replacement transistors, ICs, and other semiconductors, marketed by Philips ECG, NTE, and Thomson (SK), are available through most parts dealers (including RadioShack on special order). The ECG, NTE, and SK lines contain a few hundred parts that substitute for many thousands of others; a directory (supplied as a large book and on diskette) tells you which one to use. NTE numbers usually match ECG; SK numbers are different.

Remember that the "2S" in a Japanese type number is usually omitted; a transistor marked D945 is actually a 2SD945.

Hamfests (swap meets) and local organizations: These can be located by writing to the American Radio Relay League, Newington, CT 06111; (<http://www.arrl.org>). A hamfest is an excellent place to pick up used test equipment, older parts, and other items at bargain prices, as well as to meet your fellow electronics enthusiasts—both amateur and professional.

A We're glad to hear from you again and glad to be of service to you at precise 50-year intervals. We'll expect to hear from you again in 2048.

Seriously, for readers who may not know it, **Electronics Now** is one of the oldest electronics magazines in the business; it was formerly **Radio-Electronics**, and before that, **Radio Craft**. Our founder, Hugo Gernsback, was a prominent advocate of new technology, and was also a pioneer science fiction publisher.

On videotape erasing, see this col-

umn, August 1998, page 11. The key is going to be getting the tape close enough to an AC electromagnet. You might modify a video tape rewinder to spool the tape past a bulk eraser. Driving the electromagnet with a frequency higher than 60 Hz might also help. Apparently, videotape erasing is surprisingly difficult; we'd like to hear from readers who have built tape erasers that work well.

Have LCD, Want To Use It

Q I have a back-lighted 10.5-inch-diagonal LCD screen, a Sharp LM64C35P, from a deceased notebook computer. Can I connect it to a desktop computer as a monitor? Is a pinout diagram available? — B. M., U.S. Navy

A You can get detailed information about Sharp display panels from <http://www.sharpmeg.com> or by contacting Sharp Microelectronics, 5700 N.W. Pacific Rim Blvd. M/S 20, Carnas, Washington 98607. Interfacing the panel to a PC sounds like a challenging project; as you surmised, it doesn't take the video signal from an ordinary VGA card, but instead responds to digital commands from the computer.

MM5369 Substitute Found

In August (pp. 12-13) we lamented National Semiconductor's decision to discontinue the MM5369AA/N IC, which produces a precise 60-Hz square wave from a 3.58-MHz color TV crystal.

Figure 3 shows a substitute. Now that 4.9152-MHz microprocessor crystals are a standard item, all you have to do is divide that frequency by 16,384 (= 2¹⁴) and then by 5 to get 60 Hz. That means you can build a crystal-controlled source of 60 Hz with two chips, and if you want 50 Hz all you have to do is change one connection. The output waveform does not have a 50% duty cycle, but that's no problem if what you're doing is controlling a clock. To get precise frequency control, make one of the capacitors C1 or C2 variable. You can use 74HC chips (74HC4060, 74HC4017) if the supply voltage is 5 volts.

Plated-Through Holes

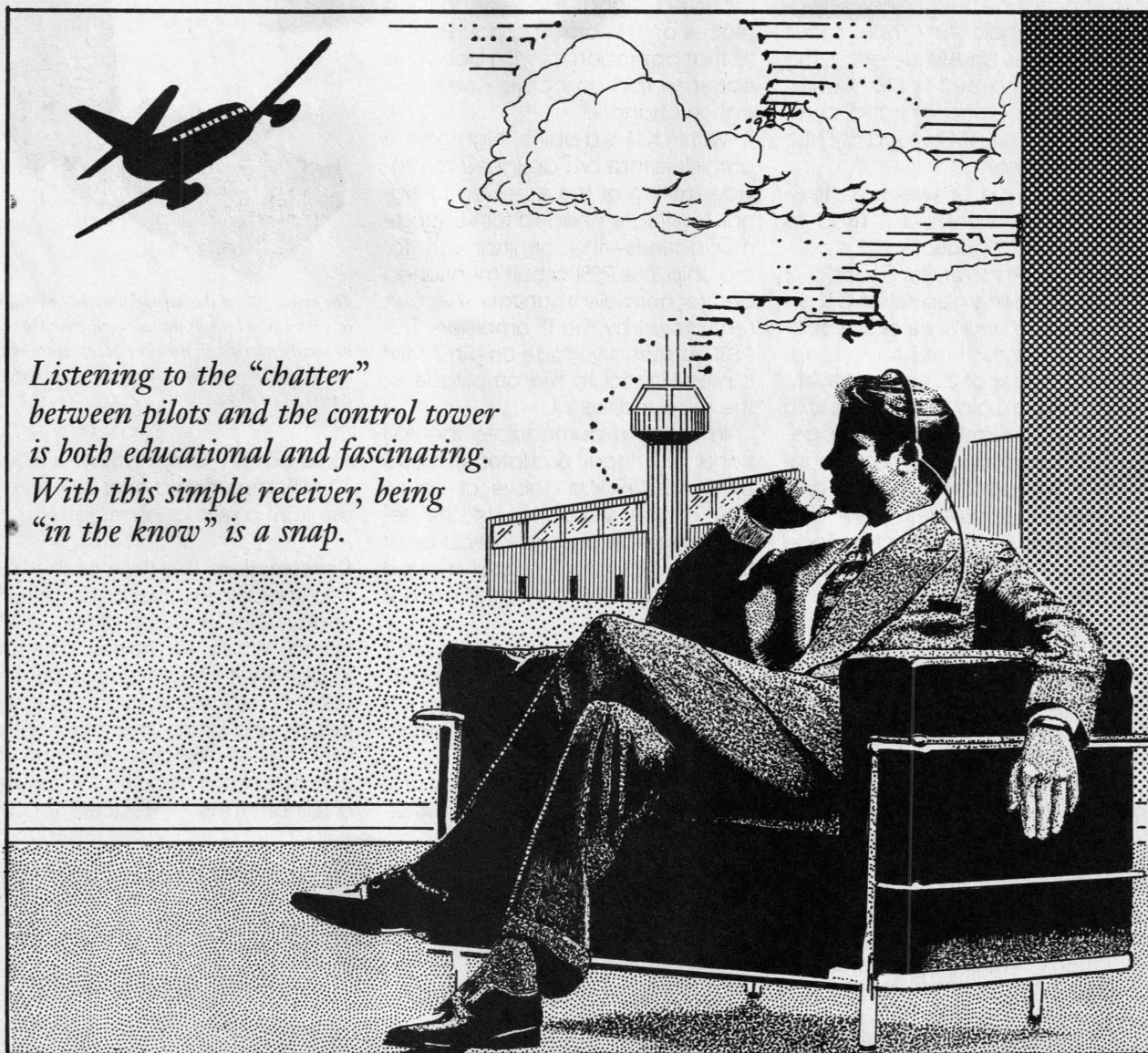
In the April issue, reader E. V., of Toledo, Ohio, asked how to make plat-

Erasing Videotapes

Q The last letter I wrote to your predecessor magazine with a question was in 1948. You were most helpful. I have another question I hope you can help me with.

Do you have any ideas on construction of a bulk eraser for video tapes? I don't want to wear out my VCR by running each tape through it just to erase. My collection of your magazines goes way back before the invention of VCRs but I can't seem to locate a project of this type. — J. L. A., Mesa, AZ

THE AIRPORT BUDDY



Listening to the "chatter" between pilots and the control tower is both educational and fascinating. With this simple receiver, being "in the know" is a snap.

If you find yourself spending a lot of time at airports waiting for an outgoing or incoming flight, or if you have an interest in aviation in general, you've probably wondered about the type of communication that occurs between the pilots and the tower. Being "on the inside" has always fascinated people. If you want to be "in the know" in terms of aviation operations, then the Airport Buddy is a must-build project. This low-cost receiver will let you eavesdrop on communications between the control tower and aircraft flying above, and you will be able to keep up with what's going on. Even if you don't have that type

ANTHONY J. CARISTI

of interest in aviation radio, the Airport Buddy will make a unique gift for a friend or relative who does fly frequently.

The Airport Buddy is a one-chip VHF superheterodyne receiver that is easy to build, compact, and small enough to fit into a pocket. It is designed to receive audio communications in the 108- to 135-MHz frequency band, which is used for ground-to-air communications.

The project consists of a small PC board that is powered by a common 9-volt battery and fits in a small case. A pair of headphones is used

to hear the received audio signals. A knob-adjustable frequency control (either a single- or a multi-turn type) is used to tune in whatever conversations are taking place at the time. An optional volume control is easily added to the circuit.

Since all of the difficult design requirements of the circuit have been taken care of by a suitable PC board layout, a receiver that works well is easily built simply by following the construction details given here.

How It Works. The schematic diagram for the Airport Buddy is shown in Fig. 1. The heart of the receiver is IC1, a sophisticated IF (intermediate

frequency) amplifier chip that contains a mixer, a local-oscillator transistor, and a pair of high-gain IF amplifiers. Also included is a received-signal-strength indicator (RSSI) circuit that is used as an AM detector. The chip also has a built-in FM-quadrature detector; since air-traffic communications use AM instead of FM, it is not used here.

A short piece of wire or a telescoping radio antenna is used to pick up the RF signals. The RF is coupled to the mixer input of IC1 through coupling capacitor C1.

The emitter and base of the built-in local-oscillator transistor appear at pins 3 and 4 of IC1, respectively. The circuit is a Colpitts oscillator, with C3 and C4 forming the voltage-divider capacitor pair that is connected between the ground and the internal transistor's base and emitter. Coil L1 and varactor diode D2 form a parallel-tuned circuit that sets the local-oscillator frequency. Potentiometer R6 is a front-panel tuning control that adjusts the DC bias on D2. That sets the local-oscillator frequency for tuning the radio; it can be varied over the range of about 108 to 135 MHz.

The intermediate frequency of the circuit, 455 kHz, is set by RES1

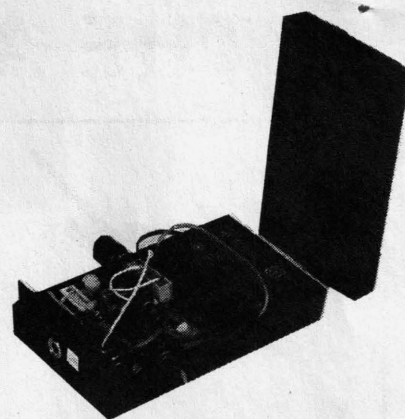
and RES2. An interesting feature of the Airport Buddy is the lack of RF preselection. That feature lets a received signal appear in two places on the dial. The advantage to that approach is that it becomes easier to find an active communication channel.

Within IC1 is a pair of high-gain IF amplifiers that are designed to provide limiting of the received RF signal. Limiting is needed for FM communications—the original use for the chip. The RSSI circuit mentioned before normally monitors the current drawn by the IF amplifiers. The RSSI outputs a voltage on pin 7 that is proportional to the amplitude of the received signal.

In order to demodulate the AM signal, the local oscillator is tuned either to 455 kHz above or below one of the sidebands of the received signal frequency. Because of that, the RSSI circuit makes a good demodulator for the received AM signals.

The demodulated audio on pin 7 of IC1 is coupled to the gate of Q1 through C8. The amplified signal is then coupled to an external headset through T1.

Power to operate the circuit is provided by a common 9-volt transistor-



The small size of the Airport Buddy's PC board lets the entire unit fit into a small case that can be easily carried in a shirt pocket. Here is the author's prototype. Note that ANT1 is simply a length of insulated wire.

radio battery. Current draw is about 10 milliamperes, giving over 35 hours of use from a new alkaline battery.

Construction. Because of the high frequencies involved, the Airport Buddy should be built on a PC board. When RF frequencies are involved with any circuit, a double-sided board with a large solid-copper ground plane helps to contain any stray RF radiation from the circuit. Any circuit trace that is carrying a signal in the RF-frequency band

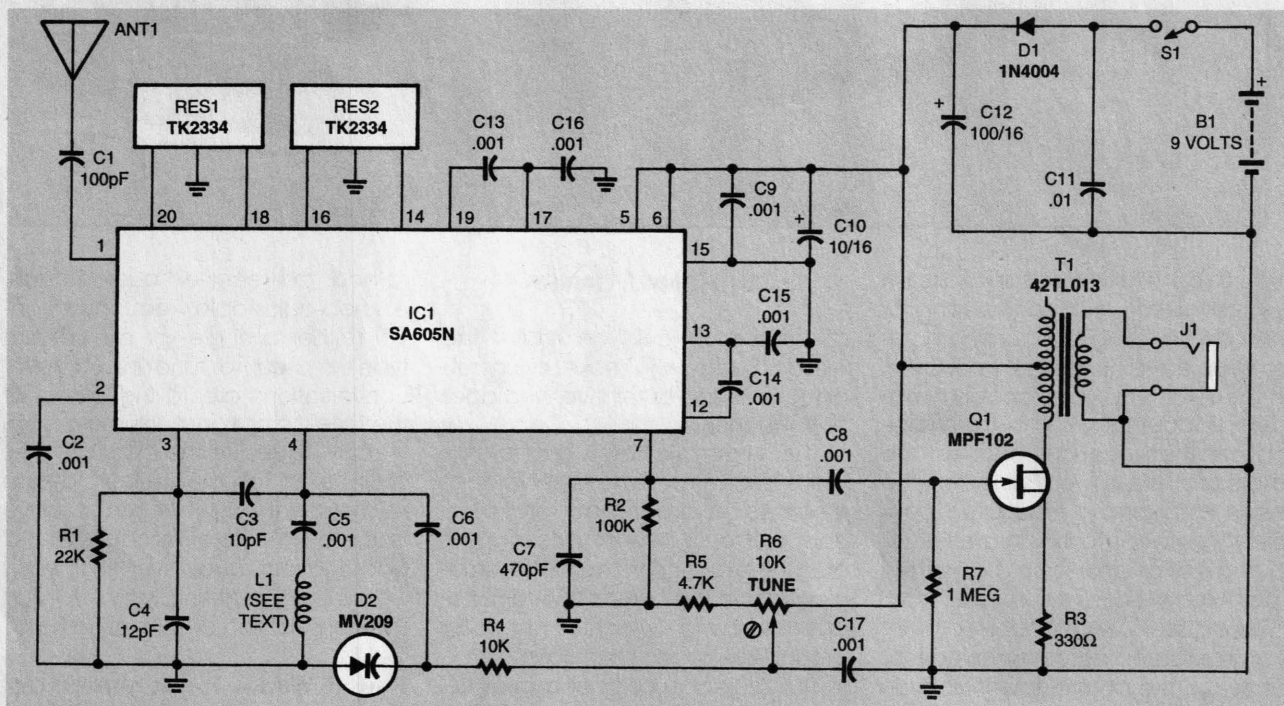


Fig. 1. The Airport Buddy is a simple receiver that is tuned to pick up frequencies used by aircraft. Although IC1 was originally designed to be used in an FM radio, it does a superb job with the AM-style method used by aircraft and airports.

**TABLE 1
COMPONENTS GROUNDED
TO GROUND PLANE**

R1	C11
R2	C15
R3	C16
R5	C17
C2	IC1 pin 15
C4	D2
C7	L1

can act like a miniature broadcast antenna, resulting in the disruption of any other receivers that happen to be close enough to the source of the RF leakage.

Foil patterns for the Airport Buddy have been included. If you do not wish to fabricate your own board, one may be obtained from the source given in the Parts List. If you are going to etch your own board, note that only one foil pattern is needed. That pattern is for the solder side of the board—the component side is left as a single ground plane. An easy way to etch the board is to coat the ground-plane side with resist; common shellac can be used. The other "foil pattern" only shows the locations of the holes that should be cleared of copper ground plane before building the circuit. Use a sharp knife to clear a small amount of copper from around the indicated holes. The component leads that will be inserted into those holes should not touch the ground plane. The other holes that will not be cleared will be used as ground connections on the top side of the board.

The parts-placement diagram shown in Fig. 2 is to be used for locating the various components if you are using a purchased board or one made from the foil pattern. Be sure to double-check the orientation of the polarized components before soldering them in—especially the semiconductors. For example, D2 has an unusual package—it looks like a two-leaded transistor. Note, however, that RES1 and RES2 are *not* polarized. Do not install IC1 or C9 at this time.

Because of the high frequencies involved, keep the lead lengths as short as possible. Before installing T1, use an ohmmeter to measure the resistance of each of the windings. One winding, the primary, will

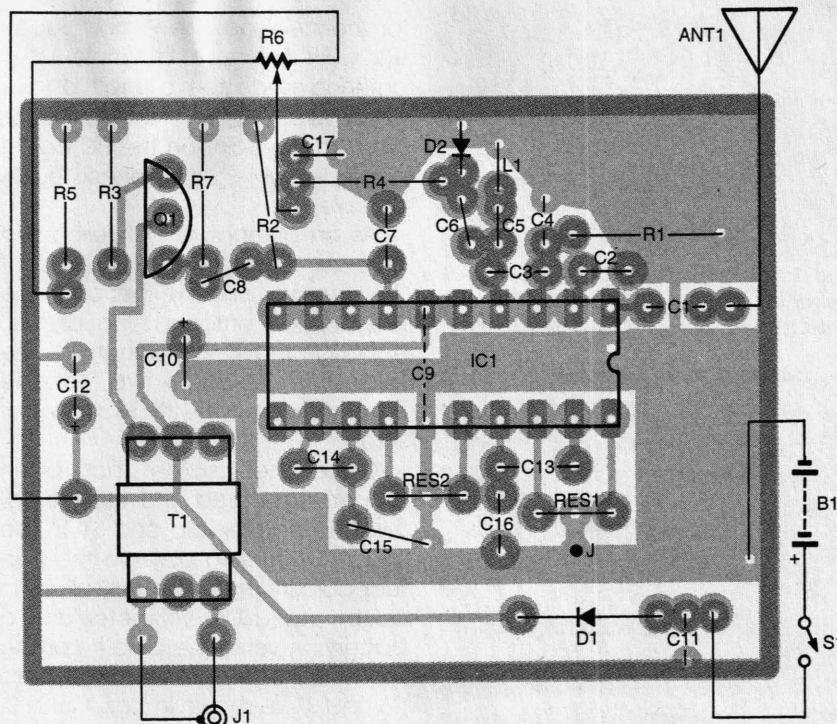


Fig. 2. Here's the parts-placement diagram for the Airport Buddy. Note that C9 is tack-soldered to the solder side of the board between the pins of IC1. Keep the body of C9 flat to the board.

measure about 50 ohms; the secondary will measure about 1 ohm. Mark the windings with a "P" and an "S" to be sure that the transformer is inserted properly into the circuit. The primary of the transformer is driven by Q1; the secondary feeds the earphone jack.

Coil L1 will be made from a piece of solid wire wrapped around a form. Following the dimensions shown in Fig. 3, wrap 3½ closely-spaced turns of insulated 22- or 24-gauge wire around an 1/8-inch-diameter form such as a drill bit. Following the height dimension shown in Fig. 3, insert the coil into the board and solder it in place. The coil can also be made from bare wire; however, the turns should not short out to each other or to the ground plane.

Table 1 lists the components that must have ground connections to the ground plane on the topside of the board as well as the bottom. Additionally, one through-hole wire is used to connect the ground plane to circuit ground; it is located next to RES1.

Once those components have been installed, use an ohmmeter to check the connections of all of the ungrounded components to be

sure that there are no short circuits to ground. Carefully examine all of the solder joints—they should be shiny and smooth. Correct any problems now before proceeding.

Check that all of the holes for IC1 on the topside of the board are clear of copper except for pin 15. When IC1 is inserted, none of those

**TABLE 2
NORMAL VOLTAGES ON IC1**

Pin No.	Voltage
1	1.4
2	1.4
3	5.4
4	6.2
5	6.3
6	6.3
7	0.25*
8	2.1
9	2.1
10	3.7
11	1.7
12	1.7
13	1.7
14	1.7
15	0 (gnd)
16	1.7
17	1.7
18	1.7
19	1.7
20	5.1

* The voltage on pin 7 depends upon signal strength.

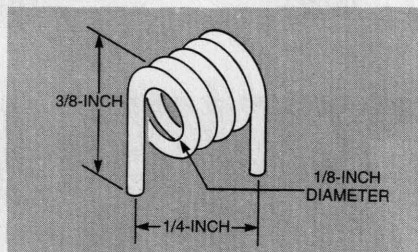


Fig. 3. It is easy to wind L1. Take a 3/2-inch length of 22- or 24-gauge insulated wire and wind 3 1/2 turns around a 1/8-inch-diameter drill.

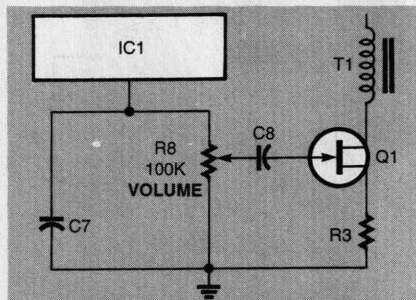


Fig. 4. If you want to add a volume control to the Airport Buddy, simply replace R2 with R8 as shown here. Lift the lead of C8 that would normally be connected to R2 and C7 and connect it instead to the wiper of R8.

pins should be touching the ground plane. Integrated circuits usually have pins designed to sit on top of the hole. If everything looks good, install and solder IC1 into the circuit. Be sure to solder pin 15 on both sides of the board.

Once IC1 has been soldered in place, turn the board over to the solder side. Capacitor C9 is tack-soldered underneath IC1 between pins 6 and 15. The leads of C9 must be cut as short as possible, and the body of the capacitor should lay as flat as possible against the board.

Any small case that can hold the PC board, a 9-volt battery, and the controls will work well. For example, a small case with a built-in battery compartment that fits easily into a shirt pocket can be used. Drill appropriate holes for J1, S1, and R6. You can use a single-turn potentiometer for R6 and draw a dial on the case, but it will be a bit more difficult to tune the Airport Buddy to a particular frequency than if you were to use a multi-turn potentiometer.

An option to consider is that of a volume control. If you want to use that option, which is detailed in Fig. 4, R8 must also be mounted in the case.

An additional hole will be needed

for the antenna. The Airport Buddy will work fine with a simple whip antenna made from a length of insulated wire. The hole should be as close to the point on the PC board where the antenna is connected as possible.

As an alternative, a telescoping antenna can be used instead of the length of wire. In either case, the length of the antenna is not critical. Although the 1/4-wavelength at the frequencies is about two feet, the receiver will work with 16 inches of wire.

Once the cabinet has been properly prepared and assembled, wire the battery clip and S1 to the circuit. Be sure to observe the correct polarity of the battery clip. If in doubt, use a DC voltmeter and a battery to verify the polarity of the leads. Make the final connections to R6 and J1 as shown in Figs. 1 and 2. Be sure to wire R6 so that clockwise rotation causes the voltage at its wiper to increase. That will give higher received frequencies with clockwise rotation of the knob. If you are going to use a stereo jack for J1, you should connect both left and right together on the jack.

Checkout Procedure. Before connecting B1 to the clip, check all of the wiring thoroughly once more to be sure that it is 100% correct. It is far easier to correct any problems or mistakes now rather than later if you find that the Airport Buddy does not work. Using an ohmmeter, measure the resistance across C12 to verify that there is no short circuit to ground on the main power line. Turn S1 off and clip a fresh alkaline 9-volt battery to the battery connector. Insert a headphone set into the jack. The Airport Buddy will work with any stereo or monaural headphones. However, the volume of the audio is directly proportional to the quality of the headset. Very low-cost headphones are relatively inefficient and will not produce as much volume as higher-quality units. Good headsets are readily available for less than \$20.00.

Turn S1 on. You should be able to hear a very soft "white noise" or static as the tuning knob is rotated over its range. If there are any aircraft nearby, you might hear voice

PARTS LIST FOR THE AIRPORT BUDDY

SEMICONDUCTORS

IC1—SA605N intermediate-frequency decoder/demodulator, integrated circuit
D1—1N4004 silicon diode
D2—MV209 varactor diode
Q1—MPF102 field-effect transistor

RESISTORS

(All resistors are 1/4-watt, 5% units, unless otherwise noted.)

R1—22,000-ohm
R2—100,000-ohm
R3—330-ohm
R4—10,000-ohm
R5—4700-ohm
R6—10,000-ohm, linear-taper potentiometer, panel mount
R7—1-megohm
R8—100,000-ohm linear- or audio-taper potentiometer, panel mount (optional volume control, see text)

CAPACITORS

C1—100-pF, ceramic-disc
C2, C5, C6, C8, C9, C13—C17—0.001-μF, ceramic-disc
C3—10-pF, NPO ceramic-disc
C4—12-pF, NPO ceramic-disc
C7—470-pF, ceramic-disc
C10—10-μF, 16-WVDC, electrolytic
C11—0.01-μF, ceramic-disc
C12—100-μF, 16-WVDC, electrolytic, low-leakage

ADDITIONAL PARTS AND MATERIALS

ANT1—Whip antenna, 16–24 inches (see text)
B1—9-volt transistor radio alkaline battery
J1—Earphone jack (stereo or monaural as needed)
RES1, RES2—455-kHz ceramic filter, Digi-Key TK2334
S1—Single-pole, single-throw, toggle or slide switch
T1—1000-ohm to 8-ohm audio transformer (Mouser 42TL013 or similar)
Enclosure (PacTek K-HM-9VB or similar), panel-mount adapter for R6, 9-volt battery clip, tuning knob, headset, hardware, etc.

Note: The following items are available from: A. Caristi, 69 White Pond Road, Waldwick, NJ 07463; Etched and drilled PC board, \$15.75; IC1, \$12.50; D2, \$3.00. Please add \$5.00 for postage and handling. NJ residents add 6% sales tax.

communications.

If the receiver is working correctly, you can take your Airport Buddy to an airport and verify that you are able to listen in on aircraft communications. Otherwise, refer to the

(Continued on page 44)

Learn how crystal-controlled oscillators produce precise, stable output frequencies by building the circuits.

DAN BECKER



THE CRYSTAL-CONTROLLED oscillator has provided stable timing and frequency signals for years, and is now an integral part of products ranging from watches and computers to handheld transceivers and satellite receivers. First introduced in vacuum-tube form, most are now transistorized. The word *quartz* in advertising copy and specification sheets is the clue that they're inside. A time or frequency base derived from a resonating quartz crystal is the next best thing to a national time standard.

In the first installment of this two part series, resonators based on the piezoelectric effect were introduced. It was pointed out that a piezoelectric crystal produced electricity if it is subjected to physical stress and, conversely, the crystal is physically distorted if a voltage is impressed across its faces. Both properties are put to use in depthfinder transducers.

In fact, the first practical application of a piezoelectric crystal was as a transducer to generate and receive sounds underwater for the detection of submarines during World War I. Later, crystal loudspeakers, microphones, and phonograph pickups were developed. The first quartz crystal-controlled oscillator was introduced in the 1920's.

Synthetic or cultured quartz

is now the dominant material of both crystal resonators and filters. The desired physical and electrical properties are obtained by cutting the quartz blank according to a set of strict rules. Packaged quartz crystal resonators are available as low-cost catalog items, and if you can't find the frequency you need, you can order a custom-made resonator.

All this is background to the subject of crystal-controlled oscillators. This article covers the fundamentals of all transistorized oscillator circuits, and includes five schematics and eight tables covering crystal-controlled oscillators complete enough for you to build your own circuits for personal instruction, experimentation or to meet a specific project requirement. All of the components are low in cost and readily

available through retail stores or mail-order houses.

What is an oscillator?

An oscillator is a circuit that generates a specific frequency and maintains that frequency within limits. A transistorized inductance-capacitance (LC) oscillator depends for its operation on the resonant interchange of energy between a capacitor and inductor for its operation; a transistor amplifier supplies pulses of energy of the proper phase and magnitude to maintain oscillations.

When used in oscillator circuits, transistors become converters that change DC electrical energy from the collector power supply into AC energy in the output circuit. The amplifying characteristic of the transistor maintains the circuit oscillations.

CRYSTAL OSCILLATORS

Two conditions are necessary to sustain oscillations. First, the feedback voltage from the collector circuit must be in phase with the original excitation voltage on the base—that is, the feedback must be positive or regenerative. Second, the amount of energy fed back to the base circuits must be sufficient to compensate for the energy losses that occur in the base circuit.

It is useful to review the concept of Q before discussing oscillator operation. Q is defined as a figure of merit in a resonant system. Equal to the reactance divided by the resistance, it represents the ability of the device or network to sustain oscillations with minimum feedback. In short, the higher the Q the more efficient the resonator.

The intrinsic Q of quartz is 10 million at 1 MHz. Although the Q value for a mounted resonator crystal is reduced to levels of 20,000 to well over a million, it is still orders of magnitude better than the best LC resonator or tank circuit.

Crystal oscillator theory.

The very high Q of a crystal oscillator significantly reduces frequency drift caused by temperature and power supply voltage variations. Moreover, crystal-controlled oscillators generate less noise than conventional oscillators with LC tank circuits, so they have a purer output signal.

The simplest crystal oscillator consists of a single bipolar transistor with a simple feedback network. Figure 1-a shows a block diagram for a generic crystal oscillator. Here, an NPN amplifier is connected in three feedback circuits. Appropriate DC bias is assumed but not shown. Figure 1-b is an equivalent circuit with components $L1$ and $C1$, and $C2$ shown. All of the crystal oscillator circuits to be discussed here have the same basic topology, and include at least two capacitors and an inductor. The crystal can be considered to be part of the feedback circuit.

Capacitors $C1$ and $C2$ include residual transistor and junc-

tion capacitance. Capacitor $C2$ can be equivalent to a parallel combination of an inductor and a capacitor. This pair functions as the crystal's third overtone selector because it is capacitive only at the crystal's overtone frequency, and inductive at its fundamental. Thus, an inductor located at $C2$ prevents oscillation at the crystal's fundamental frequency.

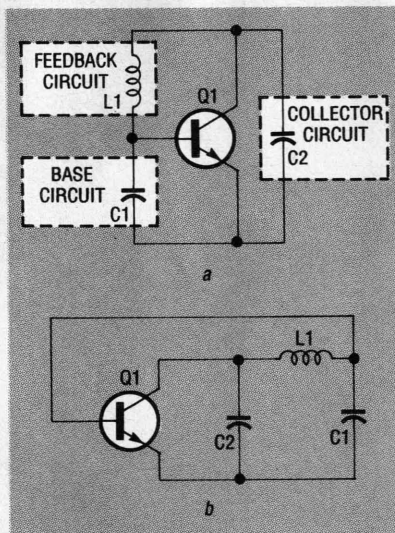


FIG. 1—BASIC OSCILLATOR CIRCUIT showing equivalent components, a and the circuit redrawn to show a feedback network, b.

In addition to amplification and feedback, an oscillator requires limiting, which occurs when an increase in the input signal no longer produces an increase in the output signal. Thus, the oscillator's output reaches a limit and stays there.

Some oscillator circuits are named for a circuit characteristic such as electron-coupled or phase-shift. However, many oscillators are named for their inventors: Among these are Butler, Colpitts, Hartley, and Miller. Five inventor-named circuit have been selected for this article; three are crystal-controlled versions of the Colpitts oscillator, one of the Pierce oscillator, and one of the Butler oscillator.

The standard Colpitts crystal-controlled oscillator has rigid load and tuning requirements, while the two semi-isolated versions are less temperamental, and are recommended as better choices for general-purpose ap-

plications. If you want very precise output frequency, the Pierce crystal oscillator is your best bet. However, if you want to experiment with oscillators, you'll find that the Butler circuit oscillates even without a crystal, you can observe the results with the crystal in or out. Table 1 compares the characteristics of each of these circuits.

Colpitts oscillators

Two versions of the Colpitts crystal-controlled oscillator are presented here: the standard and the semi-isolated. The standard circuit, shown in Fig. 2, is sensitive to variations in both crystal and load resistance. In addition, its output power is limited to less than half of the crystal's power dissipation. But, it's still a popular circuit.

Resistors $R1$, $R2$, and $R3$ DC bias transistor $Q1$. Potentiometer $R2$ allows up to about 1.5 milliamperes of emitter current. Capacitors $C3$, $C4$, and $C6$ bypass radio frequency at XTAL1's operating frequency (fundamental or overtone). Capacitor $C2$ functions like the feedback base circuit capacitor $C1$ in Fig. 1. At XTAL1's operating frequency, $L1$ and $C5$ have a net capacitive reactance, and thereby form collector circuit feedback capacitor $C2$.

For overtone crystals, $L1$ and $C5$ act like an overtone selector, preventing oscillation at the crystal's fundamental frequency. Trimmer capacitor $C1$ fine tunes feedback element $L1$. As the value of $C1$ is made smaller, the oscillator's output frequency increases.

To organize a standard Colpit-

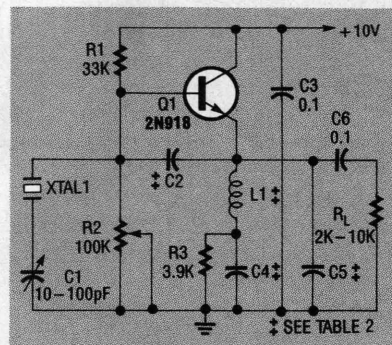


FIG. 2—A STANDARD COLPITTS crystal-controlled oscillator.

ts oscillator for your specific output frequency requirements, refer to Table 2. Note that frequencies from 1 MHz to 30 MHz are obtained with the fundamental mode, and frequencies from 35 to 60 MHz are obtained with a third-overtone crystal. The values for C2, C4, and C5 are given in picofarads, and the values for L1 are given in microhenries.

Semi-isolated Colpitts

Two versions of the semi-isolated Colpitts oscillator are described here. The first, shown in the Fig. 3 schematic, includes a fundamental-mode crystal. The second, shown in Fig. 4, is the same as that shown in Fig. 3 except that it includes overtone selector L1, C6 and radio-frequency bypass

Thus the circuit has its own built-in buffer that can drive low-impedance loads without detuning the oscillator.

However, as with the standard Colpitts circuit, the crystal is shunted by the emitter-base junction of the transistor—a low impedance. This lowers the oscillator's Q from tens of thousands to a few thousand, reducing its frequency stability. However, for most practical applications its stability is more than adequate.

Refer to the schematics shown in Figs. 3 or 4. In both circuits, resistors R1, R2, and R3 apply DC bias to transistor Q1. RF bypass capacitor C6 grounds one end of T1's primary. Capacitors C2, C5, and crystal XTAL1 form a feedback network as discussed earlier in

be the same as the crystal's frequency, use the component values given in Table 3 or Table 5. (Tables 3 through 6 contain specifications information on winding transformer T1, which is explained under the Construction section) In this case, the oscillator, the load, and T1 are all tuned to the same frequency, and each affects the tuning of the other. (The load resistor R_L should initially be a ¼-watt resistor).

If the output frequency is to be a harmonic of the crystal's frequency, use the component values given in Table 4 or table 6. As mentioned earlier, this arrangement isolates T1 and the load, enabling the circuit to work with a wide range of load impedances.

The semi-isolated Colpitts oscillator shown in Fig. 4 requires a third-overtone crystal. Therefore, L1 and C5 appear capacitive at the third overtone, but they appear inductive at the crystal's fundamental frequency; together they form the collector circuit feedback element C2 shown in Fig. 1. Capacitor C6 bypasses DC-bias resistor R3, but it is most effective at the overtone rather than the fundamental frequency.

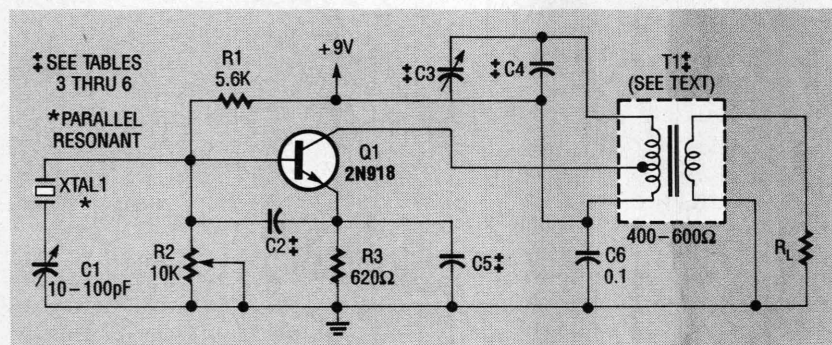


FIG. 3—A SEMI-ISOLATED COLPITTS OSCILLATOR with a fundamental-mode, parallel resonant crystal.

capacitor C3. Its operation requires a third-overtone crystal.

In both circuits, RF transformer T1 takes an output signal from Q1's collector current, but T1 is not part of the oscillator's feedback network. In addition, the output power is up to 100 times the crystal's power dissipation. Therefore, 15 milliwatts of output can be obtained with only microwatts of crystal dissipation! Moreover, if the output transformer is tuned to a harmonic of the oscillator's frequency, the RF load current is effectively isolated from Q1's fundamental RF current. Therefore, variations in the load or T1 do not affect oscillator tuning.

For example, if a 10-MHz crystal is used, the RF transformer can be tuned to 20 MHz, 30 MHz, 40 MHz, or higher MHz.

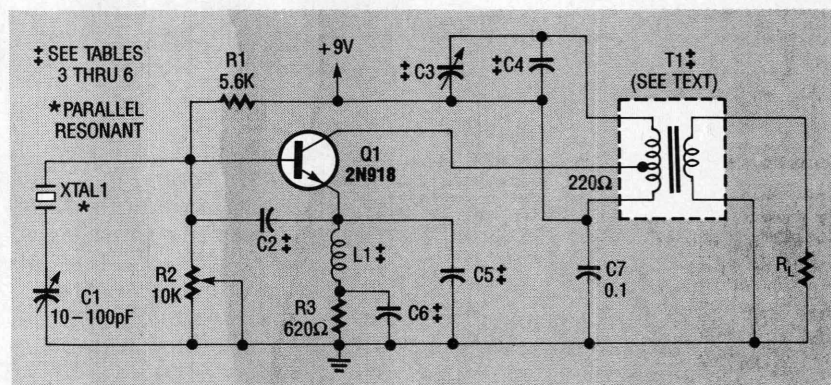


FIG. 4—A SEMI-ISOLATED COLPITTS OSCILLATOR with a parallel resonant crystal and overtone selector.

reference to Fig. 1. Trimmer capacitor C1 serves the same function as it does in the standard Colpitts circuit. Transformer T1 and trimmer capacitor C3 (with C4) form a parallel resonant tank tuned to the desired output frequency.

If the output frequency is to

Pierce oscillator.

The best feature of the Pierce crystal oscillator, shown schematically in Fig. 5, is its very high operating Q . That very high Q is maintained because the crystal is connected between Q1's base and collector (a high impedance). This os-

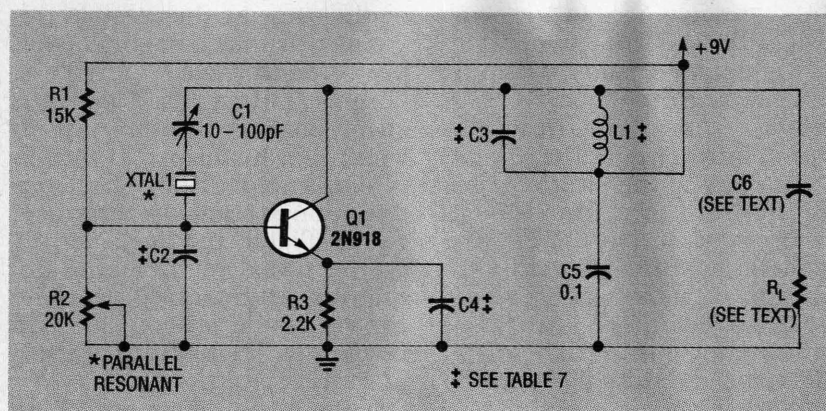


FIG. 5—A PIERCE CRYSTAL-CONTROLLED OSCILLATOR with a parallel resonant crystal.

cillator provides a very stable and accurate output frequency up to about 75 MHz. It is possible to tune this oscillator to within a few hertz of the desired frequency and expect it to remain stable there (when held at constant temperature).

If an oven-controlled crystal is used, frequency change will only be several hertz over a wide temperature range. However, the Pierce oscillator does not offer very high output power. Moreover, it requires a very high load resistance of about 3000 ohms.

Resistors R1, R2, and R3 establish the DC-emitter current. Capacitor C5 bypasses RF at all frequencies, while C4 bypasses RF current at the desired operating frequency (fundamental or, overtone only). Capacitor C2 is base circuit feedback element C1 as shown in Fig. 1, and the parallel network of C3 and L1 yields a net capacitive reactance at the operating frequency and is analogous to collector circuit feedback element C2 as shown in Fig. 1. Crystal XTAL1 forms feedback inductor L1 of Fig. 1, and as with the Colpitts oscillator, trimmer capacitor C1 fine tunes the circuit's operating frequency.

The component values of the Pierce oscillator also depend on frequency and are given in Table 7. Note that a fundamental-mode crystal permits frequencies from 1 to 25 MHz, while a third-overtone crystal is required for output frequencies from 30 MHz to 75 MHz.

If a load resistance of the

circuit operates at series resonance, making it look resistive in the circuit. It is possible to substitute a 47-ohm resistor for the crystal and tune the circuit to a wide range of frequencies with variable inductor L1. But the circuit is so sensitive to variations in load resistance that a fixed resistive load must be connected to its output.

Resistors R1, R2, and R3 set the DC-emitter current. Bypass capacitors C1, C5, and C2 place transistor Q1 in a common-base configuration, couple the collector to the load, and bypass the

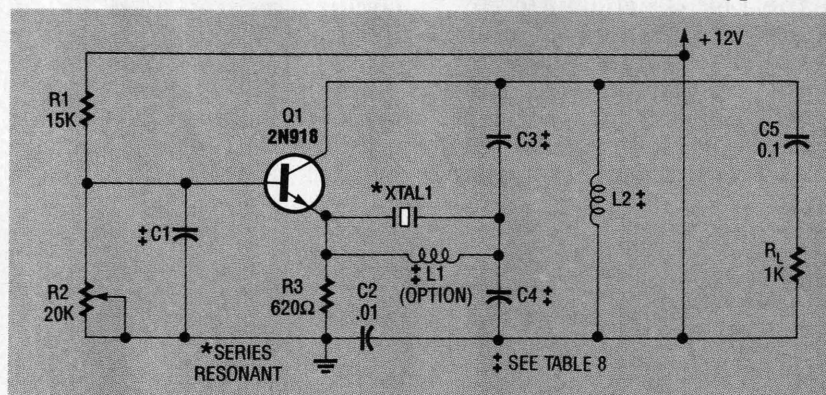


FIG. 6—A BUTLER CRYSTAL-CONTROLLED OSCILLATOR with a series-resonant crystal.

Pierce crystal oscillator is to be less than several thousand ohms, a 1- to 5-picofarad coupling capacitor must be used at C5 to prevent the low resistance of the load from detuning the circuit and preventing circuit oscillation.

Butler oscillator

A schematic of a Butler crystal oscillator is shown in Fig. 6. The Butler oscillator demonstrates what is known as *input-resistance limiting*. Transistor Q1's DC-emitter current is directly proportional to the strength of the radio frequency input signal; in addition Q1's RF-input resistance of approximately 40 ohms is inversely proportional to the DC-emitter current. Therefore, as the RF feedback increases, Q1's DC-emitter current increases, but its RF-input resistance decreases. As Q1's RF-input resistance decreases, its RF gain also decreases, and this causes the output signal strength to reach a plateau.

The Butler oscillator's crystal

positive supply lead, respectively. The feedback elements consist of capacitors C3 and C4, and inductor L2. The capacitors act like the base and collector elements C2, and C1 and inductor L2 act like the feedback circuit L1 in Fig. 1.

Crystal XTAL1 feeds some of the RF output energy back into the emitter. Because the crystal behaves like a narrow bandpass filter, the emitter current forms clean sine waves that are low in harmonics. Inductor L1 cancels the detuning effects of the crystal's static capacitance C_p . However, if you build this circuit, get it to oscillate first without L1. Then, after it is working, install and fine tune L1 to obtain a precise output frequency.

Table 8 gives the values of the components in Fig. 6 that are shown without values. Note that a third-overtone crystal is necessary to obtain output frequencies from 20 to 55 MHz, while a fifth-overtone crystal is needed to obtain output frequencies from 60 to 100 MHz.

Selecting components

Remember that many common passive electronics components such as capacitors, resistors, and inductors that perform well at audio frequencies become inefficient and lossy in the radio frequencies. For example, the parasitic inductance of a wirewound resistor can be significant in a radio-frequency circuit. Keep this in mind when selecting oscillator components.

Power-supply regulation is important for stable oscillator operation. Low-Q oscillators that are tunable over a wide range of frequencies require very stable, low-noise supplies, but that is not a requirement for high-Q crystal-controlled oscillators because they are generally immune to voltage spikes and noise. Thus, while regulation is important, it can be obtained with low-cost integrated circuit regulators.

Even with minimal regulation, it is recommended that the power supply positive voltage lead be bypassed with a 6- to 10-microfarad tantalum capacitor and a 0.01 microfarad ceramic disc capacitor. Capacitors suitable for RF bypassing and tuning should have Q values of 100 or more, and most general purpose ceramic disc capacitors meet that requirement.

The crystal oscillators described here need only one bipolar transistor. In general, any NPN transistor with a gain-bandwidth product of at least 650 MHz is suitable. Possibilities include the 2N918 (shown in all of the circuits in this article), the MPSH-10, and the 2N2857. The 2N918 is available in plastic packages and metal cans. If you use 2N918's in metal cans, be sure to ground their cases. For frequencies below 50 MHz, a 2N3904 switching transistor will perform satisfactorily.

Inductor variety

Two basic kinds of inductors (tuning coils) are generally available: air wound, and core wound. Most are wound from number 20 to 40 AWG enamel-coated magnet wire. Inductors

for RF circuits usually have both magnetic and electrostatic shielding. This is obtained by winding the coil on an iron-powder core, and then enclosing the assembly in a small metal can.

Core-wound inductors offer high Q's, good temperature stability (usually ± 200 ppm/ $^{\circ}$ C), and are small. Powdered-iron core materials are often mixed with other materials to make special powdered-iron alloys. Each is formulated to yield a high Q and optimum temperature stability over a specified frequency range. Oscillators or tuned circuits usually require Q's between 60 and 120. Before selecting a coil, examine the manufacturer's data to verify the coil's Q and usable frequency range.

Winding transformers

You can make your own radio-frequency transformer T1 for the semi-isolated Colpitts crystal oscillators shown in Figs. 3 and 4 from the toroidal cores and magnet wire specified in Tables 3 through 6. Under the column heading T1 (primary) these tables give the total number of primary turns, the wire gauge, and the designation for the appropriate core (e.g. T-80-2, T-50-2, T-50-10, T-50-17).

The first letter in this core code, T, stands for toroid, and the first number stands for the core outside diameter in fractions of an inch (e.g. 80 = 80/100 inch, 50 = 50/100 inch). The third number in the code designates a specific powdered-iron composition.

When winding wire on the core, first wind on approximately 12.5% of the total primary turns on the core (or three turns, whichever is larger). Then make a tap by twisting together several inches of wire to form a loop, and continue winding until all the primary turns have been wound. The loop can then be formed into a tap by cutting the loop and scraping the insulation off of the ends. This tap is then ready to be connected to Q1's collector as shown in Figs. 3 and 4 and sol-

dered during the circuit assembly procedure.

The secondary of transformer T1 should reflect the load impedance shown on the schematics of Figs. 3 and 4 (at the primary's tap). The turns ratio between the secondary and the primary's tap should be the square root of tap resistance divided by load resistance. The secondary winding can be wound from the same gauge wire as the primary.

For example, to reflect a tap resistance of 500 ohms, a 50-ohm load would require a primary-tap to secondary-turns ratio of 3.16:1 (the square root of 500 divided by 50). Therefore, if the primary tap consists of three turns, use a one turn secondary. After you find that the circuit oscillates with the initial transformer, you can experiment by substituting other transformers with different turns ratios.

Oscillator construction

Carefully designed printed-circuit boards are preferable to standard perforated boards as substrates for radio-frequency oscillators to minimize noise and interference. When building radio-frequency circuits, it is important that all components be inserted so they lie as close as possible to the board.

Use coaxial cable or TV twin-lead to conduct high-frequency signals for any distances over an inch. RCA-type audio connectors work well up to frequencies of 30 MHz, but BNC, F, and equivalent 50- or 75-ohm connectors should be used at the higher radio frequencies.

The Colpitts and Pierce crystal-controlled oscillator circuits discussed here include parallel-resonance crystals, but the Butler oscillator has a series-resonance crystal. In all circuits the crystal's load capacitance rating can be from 12 to 32 picofarads.

However, the higher frequencies require a smaller load capacitance so the crystal will provide enough inductive reactance to prevent oscillation above the desired frequency. Therefore, a 12- to 20-picofarad

load capacitor should be used in circuits expected to operate at frequencies above 15 MHz.

After building the standard Colpitts oscillator, and before trying it for the first time, set variable capacitor C1 to its maximum capacitance value. The following start-up directions apply to all Colpitts oscillators and the Pierce oscillator, but *not* the Butler oscillator:

- Initially, couple a 4.7 K ohm resistor to the 0.1 microfarad load-coupling capacitor (C6 in Figs. 2 and 5).

- Adjust trimmer potentiometer R2 until the circuit oscillates. When the circuit is oscillating properly, a different load resistor value can be substituted. For 0.1 microfarad load-coupling capacitors, R_L must be a resistive load of 2 to 10 K ohms

- For coupling a low-impedance load, use a 1- to 47-picofarad capacitor.

- If the output frequency is to be the same as the crystal's resonant frequency, refer to the component values given in Table 3 or Table 5. In this case, the

oscillator, the load, and T1 are all tuned to the same frequency, and each affects the tuning of the other. Therefore, a 1/4-watt resistor should be inserted initially at R_L .

After completing the Pierce oscillator, lightly couple a radio-frequency or oscilloscope probe to Q1's collector with a 5 picofarad capacitor. Then carefully adjust trimmer potentiometer R2 until the circuit oscillates properly.

As stated earlier, a 47-ohm resistor can be substituted for the crystal in building the Butler oscillator. The circuit can be tuned to a wide range of frequencies with the variable inductor L2. Be sure to keep a fixed resistive load connected to the oscillator's output because it is sensitive to variations in load resistance.

It was also stated earlier that a second optional inductor L1 will cancel the detuning effects of the crystal resonator's parallel capacitance C_P . Complete the construction and make sure the circuit oscillates before installing L1 and fine tuning it. **R-E**

RANDOM DOTS

continued from page 23

eyes slightly. If you do that, however, the image will be inverted; whatever would appear as floating above the page would now be recessed.

Where do they come from?

Our random-dot images were created on a PC with the *Stare-EO Workshop* software from N.E. Thing Enterprises (P.O. Box 1827, Cambridge, MA 02139, 617-621-7174). The software lets you turn graphics, text, and PCX files into professional-looking random-dot images. The \$40 disk for a PC-compatible computer can run on nearly any machine. All it requires is 512K of memory, and an HGA, CGA, MCGA, EGA, or VGA display. A mouse and hard disk make the program a little easier to use, but neither is required. The images you create can be printed on most graphics printers. There's also a Mac version of the software available for \$35.

continued on page 90

POWER SUPPLY

continued from page 46

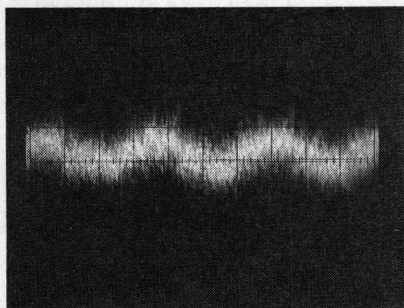


FIG. 4—THERE IS ONLY 20 MILLIVOLTS of ripple in the 250-volt DC output.

back to the voltage range and adjust the HV ADJUST knob for 250 volts. Adjust the voltmeter calibration trimmer R8 so the meter reads full scale.

Although the outputs can "float," it's best to connect one of them to the ground jack. Due to the voltage rating of typical potentiometers we do not advise floating the output more than 150 volts above ground. **R-E**

PARTS LIST

All resistors are 1/4-watt, 5%, unless otherwise noted.

R1—6.8 ohms
R2—10 ohms
R3, R4—220 ohms
R5—470 ohms, 1-watt
R6—1000-ohms, potentiometer
R7—5000-ohms, potentiometer
R8—50,000-ohms, potentiometer
R9, R10—220,000 ohms, 1/2-watt
R11—470,000 ohms
R12—1 megohm, panel-mount potentiometer

Capacitors

C1—C4—0.01 μ F, 500 volts, ceramic disk
C5—0.1 μ F, ceramic disk
C6—220 μ F, 25 volts, electrolytic
C7, C8—220 μ F, 200 volts, electrolytic

Semiconductors

IC1—LM317T variable positive regulator
D1—D5—1N4005 diode (600V, 1A)
Q1—2SC1308 NPN high-voltage

transistor, TO-3 type (Radio Shack #276-2055)

Q2—2N3904 NPN transistor

Other components

T1, T2—120/25.2 volt center-tapped 2-amp power transformer
J1—banana jack, red
J2—banana jack, black
J3—banana jack, green
F1—1/2-amp slow-blow fuse
S1, S2—DPDT switch, (6A, 250VAC)

M1—1 milliamp panel-mount DC meter

NE1—neon indicator lamp with built-in series resistor

NE2—neon lamp

Miscellaneous: Fuse holder, edge-mount perforated construction board and matching edge connector (optional, see text), enclosure, heatsink, TO-3 mounting hardware for Q1, stand-offs, knob for R12, wire, solder, etc.



Wireless Camcorder Microphone

Add professional-quality audio to your home videos with our wireless camcorder microphone—for less than \$15.00!

PAUL E. YOST

EVERY FEW YEARS, AN ELECTRONIC gadget comes along and changes our lives. Devices such as televisions, cellular phones, VCR's, and personal computers, have had a profound impact on how we live. Today's hot item is the handheld camcorder. Falling prices, smaller sizes, and greater quality have made them quite common.

If you do a lot of camcorder recording, you're sure to notice that it's always harder to record quality audio than it is to record quality video. While the camcorder might be doing a great job recording the picture, the built-in microphone might not be doing as well picking up the sound. That's particularly true for long-distance shots—remember that most lenses can zoom in on a subject, but most microphones can't.

Another difficult situation is shooting a room full of people. If there is a lot of background chatter—and there usually is—that's what the camcorder's internal mike will record. The problem is caused by the automatic audio-level control circuit

that most camcorders use—it can't differentiate between the audio you want and the babble you don't want.

The only sure-fire way to get good audio, regardless of the situation, is to use an external microphone, especially if your subject is at a distance. The most convenient mike is a wireless mike, which is the focus of this article. A good wireless mike will guarantee quality audio every time, regardless of the distance to the subject. A wireless mike also eliminates recording those annoying camcorder noises such as the sound of the autofocus mechanism.

While camcorder prices have fallen, microphone prices have not. Unfortunately, a high price does not always guarantee a high-quality wireless mike. The microphone we'll present here can be built from readily available, inexpensive parts. It takes less than 30 minutes to assemble, and requires no complicated setup or adjustment procedures. You don't even need any complicated test equipment. The resulting quality will surprise you.

Operation

The wireless microphone is basically a short-range, low-power FM transmitter. The circuit, as shown in the block diagram of Fig. 1, consists of three main sections: the microphone element, the audio amplifier, and the RF oscillator. Figure 2 shows the schematic.

The heart of the circuit is the oscillator section, which is built around a 2N3904 transistor (Q1). A parallel-resonant LC, or "tank" circuit, consisting of L1, C2, and D2, determines the operating frequency.

Varactor diode D2, a voltage-variable capacitor, tunes the circuit. To understand how the varactor works, recall that a capacitor is basically two conductors separated by an insulator. A reverse-biased diode is similar to a capacitor in that it has two electrodes (the anode and the cathode) separated by an insulator (the reverse-biased junction). Consequently, a reverse-biased diode acts like a capacitor. Although all diodes exhibit that effect, a varactor diode is designed to have as much capacitance as possible.

An oscillator is basically an

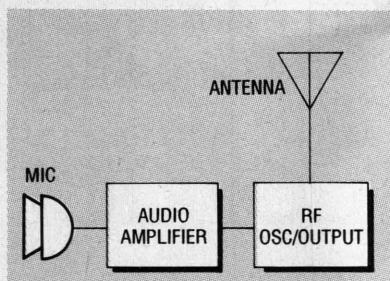


FIG. 1—BLOCK DIAGRAM. The circuit consists of the microphone element, the audio amplifier, and the RF oscillator.

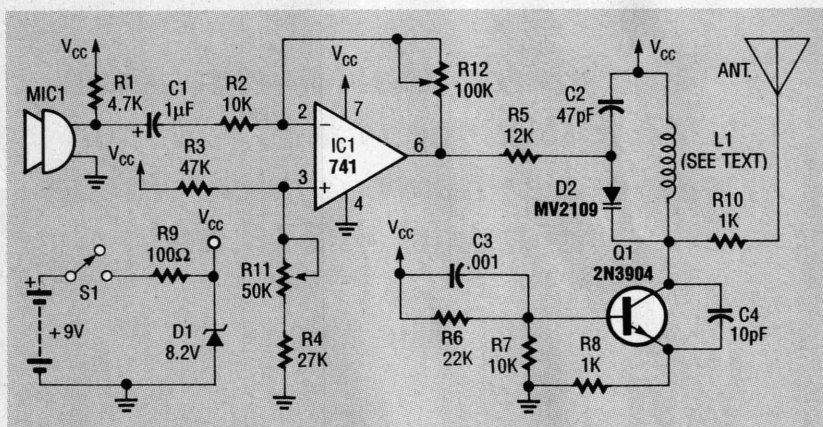


FIG. 2—The wireless microphone is a short-range, low-power FM transmitter.

amplifier with positive feedback. In our circuit, feedback is provided by capacitor C4, so that the part of the oscillation generated in the tank circuit is coupled back from Q1's collector to its emitter. Resistors R6 and R7 bias Q1. Capacitor C3 is the base bypass and R8 is the emitter load. If you want to see the oscillation, you can connect an oscilloscope to the top of R8 with a $\times 10$ probe. The initial oscillation should be somewhere between 60 and 110 MHz, so a high-frequency scope must be used. If an oscilloscope is unavailable, you can use a frequency counter to check the signal.

The audio amplifier is a 741 op-amp chosen because it is one of the most common and versatile op-amps. It is wired as an inverting amplifier with a variable negative feedback gain control. The purpose of the op-amp is to amplify the microphone signal, and to electronically tune the oscillator frequency.

An op-amp normally requires a bipolar power supply. Fortunately, op-amp circuits can be made to operate from a single

supply by placing a DC offset voltage on one of the inputs.

An op-amp is basically a differential amplifier with two inputs and one output. One input is inverting while the other is non-inverting. A signal applied to the inverting input will be phase-shifted (or inverted) 180 degrees at the output. A signal applied to the non-inverting input will remain unchanged in phase at the output. If a signal

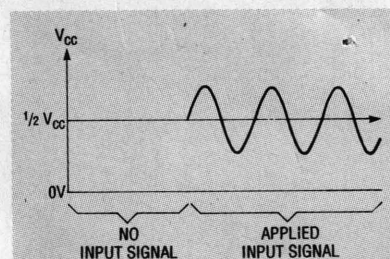


FIG. 5—WITH THE UNUSED INPUT referenced to one-half of V_{CC} , the output signal can swing both positive and negative around the new reference.

Therefore, the difference between the two inputs is the incoming signal, which is amplified and passed through to the output. This circuit works fine when both negative and positive supplies are used. However, if only a positive supply is used, the output signal would only be able to swing positive, as shown in Fig. 4.

For the op-amp to work properly from a single supply voltage, the circuit must be able to produce both positive and negative signal swings. The easiest way to do that is to offset the output reference above ground. That's accomplished by referencing the unused input to one-half of V_{CC} , instead of connecting it to ground. In our circuit, that's done by the voltage-divider network of R3, R4, and R11. The output signal can swing both positive and negative, except that it does so around the new reference as shown in Fig. 5.

Normally, the DC component of the output is removed before the signal is passed on to the next stage, usually by a coupling capacitor. In this project, however, we do not remove the DC offset. Instead, it is used later to tune the oscillator's center frequency.

The cathode of tuning element (varactor D2) is connected to V_{CC} via coil L1, and its anode is connected to the op-amp output via R5. Since the cathode is at V_{CC} and the anode is at approximately one-half V_{CC} , the varactor is reverse-biased, which is its normal operating condition.

One of the primary factors that determines a capacitor's value is the distance between its

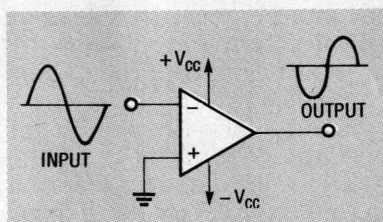


FIG. 3—NORMALLY ONLY ONE OP-AMP INPUT is used for the signal and the other is referenced to ground.

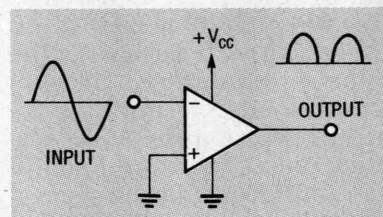


FIG. 4—IF ONLY A POSITIVE SUPPLY IS USED, the output signal can only swing positive.

is applied to both inputs simultaneously, their difference appears at the output, multiplied by the gain of the circuit.

Normally only one op-amp input is used for the signal and the other is referenced to ground as shown in Fig. 3. Since the unused input is referenced to ground, it effectively stays at a zero-volt potential.

plates. By controlling the amount of reverse bias applied to the varactor, we can control the thickness of the varactor's barrier region, and consequently the distance between its "plates." When you increase or decrease the reverse bias, the junction barrier increases or decreases. This, in turn, decreases or increases the effective capacitance, and raises or lowers the frequency of oscillation.

As mentioned earlier, the reverse bias is provided by the amplifier's DC offset, created by the voltage divider on pin 3 of IC1. Note that potentiometer R11 is part of that network. When you vary R11, the offset voltage varies as well. That, in turn, changes the bias on the varactor, which results in a frequency shift of oscillation. That's how the tuning is accomplished.

The other potentiometer, R12, is used to set the amplifier's gain. As you increase R12's resistance, you decrease the amount of negative feedback, which then increases the signal gain. The audio output rides on the DC offset, and the offset provides the tuning bias. Therefore, as the signal varies, so, too, does the bias and the oscillator frequency. That's how the frequency modulation (FM) is produced.

In FM, the frequency of the carrier varies with the frequency of the modulating signal; the amount of variation or deviation is determined by the amplitude of the modulating signal. In our case, the modulating signal is the audio picked up by the microphone, so when you adjust R12 you are adjusting the modulating signal's amplitude, which increases its deviation.

A standard FM broadcast signal has a deviation bandwidth (or carrier swing) of plus or minus 75 kHz. That amount of deviation is considered to be 100% modulation. For our transmitter to work properly, we'll need to adjust it to provide approximately the same amount of deviation.

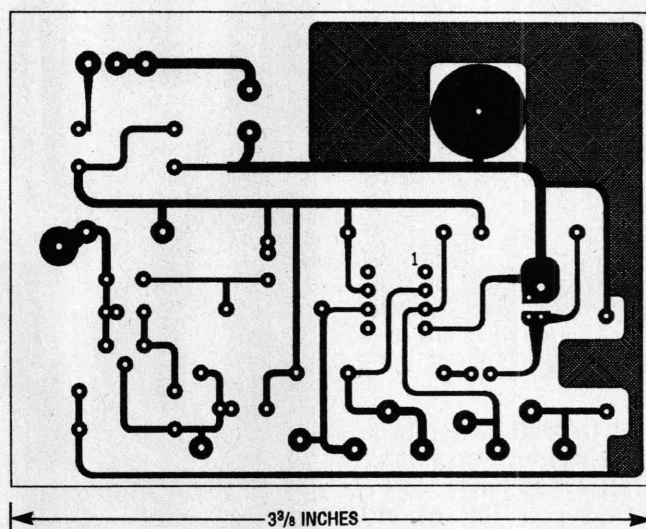
The FM transmitter uses an electret microphone. An electret is a permanently polarized piece

of dielectric material, usually a ceramic compound. It is formed by heating the material and then letting it cool in a strong electric field. That realigns the molecular structure of the material so that it retains a mild electric field after cooling. (That's analogous to the way iron can be made into a permanent magnet.) The electret is then used as the diaphragm or moving part of the microphone. The result is a small, though very sensitive, high-fidelity microphone. The microphone also contains an internal FET amplifier that adds to the microphone's fidelity and sensitivity.

Although the wireless microphone is battery operated, it uses a Zener diode voltage regulator which is absolutely essential for proper operation. A fresh battery may start out at 9 volts but, over time, the voltage slowly drops. Because the oscillator is voltage-tuned, it is voltage-sensitive. Any change in V_{CC} will shift the oscillator

transmitter, you should build it on a PC board. Point-to-point wiring could cause problems at VHF frequencies because of too much interconnection parasitic capacitance. A foil pattern has been provided so you can make your own PC board. You can also purchase a pre-made board from the source mentioned in the Parts List, as well as a kit that includes a PC board.

Install all the parts on the board as shown in parts-placement diagram of Fig. 6. There is room on the board to mount the 9-volt battery holder included in the kit if you intend to install the board in a large enough case. A metal case, like the one used for the author's prototype (see Fig. 7), will provide the best shielding for the circuit. The prototype's case is also available from the source mentioned in the Parts List. If you use the metal case, the PC board must be cut between the two "CUT A" points indicated in Fig. 6. Also, the potentiometers must be



FOIL PATTERN for the wireless microphone PC board.

frequency. This is so critical that even a 0.1-volt change can shift the oscillator by 100 kHz or more. Obviously, we want the transmitter to stay on frequency so we must make sure that V_{CC} doesn't change; the Zener diode guarantees this.

Construction

Despite the simplicity of the

mounted on the underside of the board with the battery holder secured to the bottom of the case and the PC board mounted on 7/8-inch spacers (see Fig. 8).

Even though a metal case provides superior shielding, we packaged the board in the pocket-sized plastic case, and the transmitter seems to work just fine. However, to get the

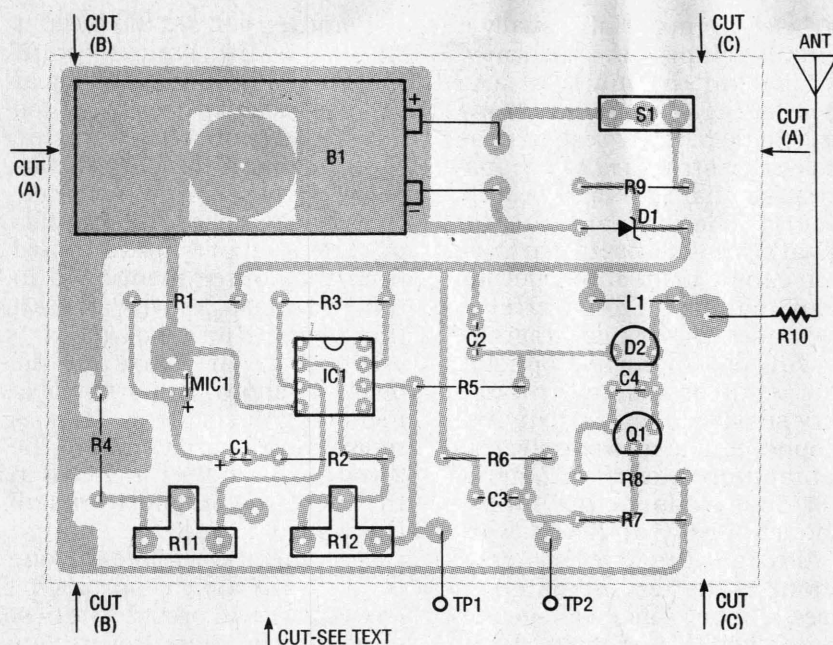


FIG. 6—PARTS-PLACEMENT DIAGRAM. There is room on the board for the 9-volt battery holder if you are installing the board in a large enough case.

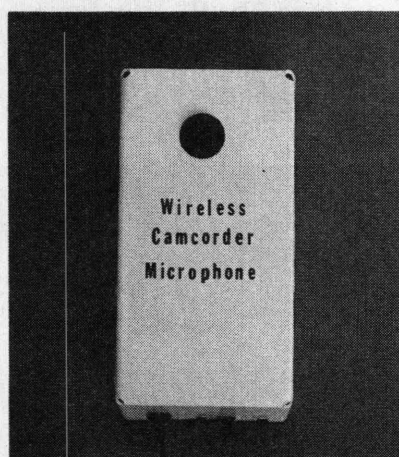


FIG. 7—A METAL CASE, like this one, provides the best possible shielding for the circuit.

board to fit in that case, three cuts must be made between the "CUT A," "CUT B," and "CUT C" points indicated in Fig. 6, and R10, which connects to the antenna, must be tack-soldered to the end of L1. Also, the potentiometer leads must be trimmed as shown in Fig. 9 to decrease their overall height so that they fit in the case. The plastic case, made by Pac-Tec, has a built-in compartment for a 9-volt battery. Figure 10 shows how the board fits in the Pac-Tec case.

The final component to connect to the board is the antenna. A 30-inch length of wire is recommended because that's

isolation for the oscillator, greatly enhancing its stability. Unfortunately the 1K resistor does decrease the unit's range. Without it, the signal will travel about 300 feet. With it, the range is approximately 100 feet, which is comparable to most commercially available products, and adequate for most video production work.

Testing

Unless you have a defective part or have misplaced something on the board, the transmitter should work as soon as power is applied. The easiest way to test it is to tune a standard FM radio to an unused frequency, and adjust R11 until you hear your voice coming from the radio's speaker. It's usually easier to adjust R11 until the microphone can be heard

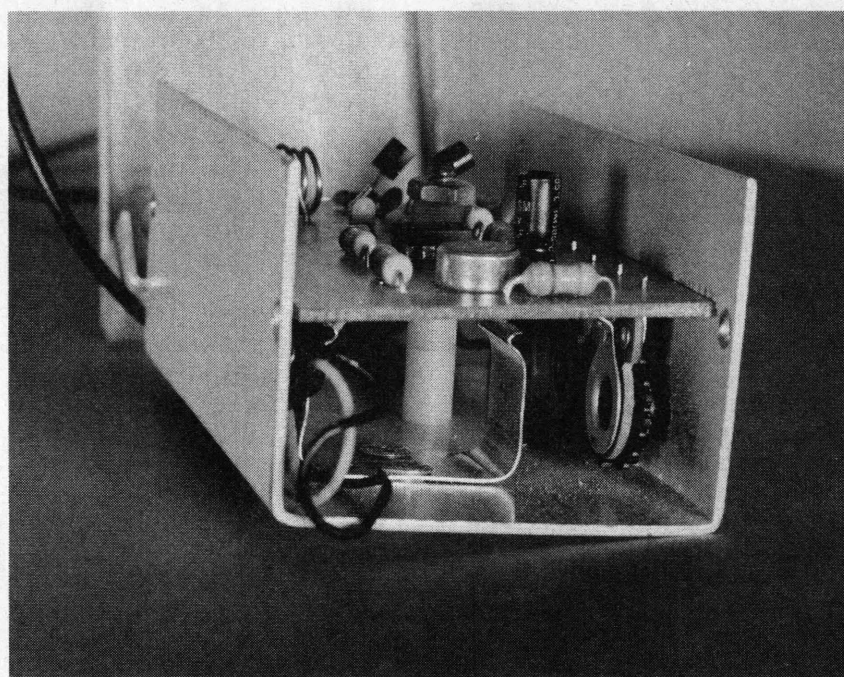


FIG. 8—TO USE THE METAL CASE, the PC board must be cut between the two "CUT A" points indicated in Fig. 6, and the potentiometers must be mounted on the underside of the board.

the quarter-wavelength of 98 MHz (the center of the FM band). But you are certainly free to use whatever length antenna works best for your needs, such as the 12-inch telescopic antenna we used for the plastic case.

The antenna connects to the collector of Q1 via a R10, a 1K resistor. The resistor provides

over the radio, and then fine tune the signal by adjusting the receiver.

You will probably have to adjust R12 as well for the proper volume level. Simply listen to the signal on an FM radio; if R12 is set too low, the audio will sound weak. If R12 is set too high, the audio will be too loud

PARTS LIST

All resistors are 1/4-watt, 5%.

R1—4700 ohms
 R2, R7—10,000 ohms
 R3—47,000 ohms
 R4—27,000 ohms
 R5—12,000 ohms
 R6—22,000 ohms
 R8, R10—1000 ohms
 R9—100 ohms
 R11—50,000 ohms, potentiometer
 R12—100,000 ohms, potentiometer

Capacitors

C1—1 μ F, 25 volts, electrolytic
 C2—47 pF, 25 volts, ceramic
 C3—0.001 μ F, 25 volts, ceramic
 C4—10 pF, 25 volts, ceramic

Semiconductors

IC1—741 op-amp
 D1—8.2-volt Zener, 1/2-watt
 D2—MV2109 varactor diode
 Q1—2N3904 NPN transistor, or equivalent
 S1—SPST switch
 MIC1—electret microphone

Other components

L1—2.5 turns of #18 wire on a 5/16-inch diameter form.

Miscellaneous: 9-volt battery and connector, battery holder, project case, PC board, 30 inches of antenna wire, solder, etc.

Note: The following items are available from Paul E. Yost, P.O. Box 32291, Louisville, KY 40232:

- A kit of parts including the PC board (no case)—\$14.95 plus \$1.50 S&H
- PC board only—\$6.95 (postpaid USA)
- Metal project case (drilled and with rubber grommets)—\$6.95 plus \$1.75 S&H

Kentucky residents must please add 6% sales tax.

- The Pac-Tec HML-9VB plastic case sells for about \$5. Call Pac-Tec at (800) 220-9800 for a distributor nearest you.

or distorted.

If you have any trouble, the following steps should help you locate the cause and solve it:

1. Check for 8.2 volts DC at the cathode of D2. If a voltage is missing or low, you either have a bad battery, a defective Zener diode, an open R9, or a short circuit on the board.
2. Check for 8.2 volts DC at the transistor collector, pin 7 of IC1,

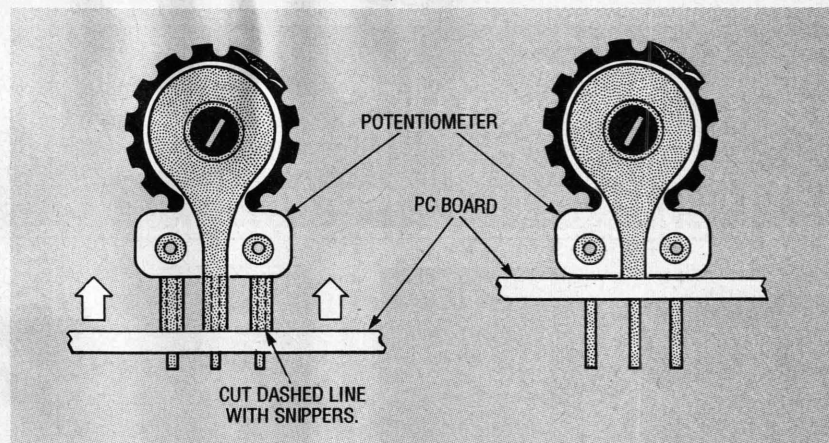


FIG. 9—TO USE THE PLASTIC CASE, the potentiometer leads must be trimmed as indicated by the dashed lines to decrease their overall height so that they have as low a profile as possible.

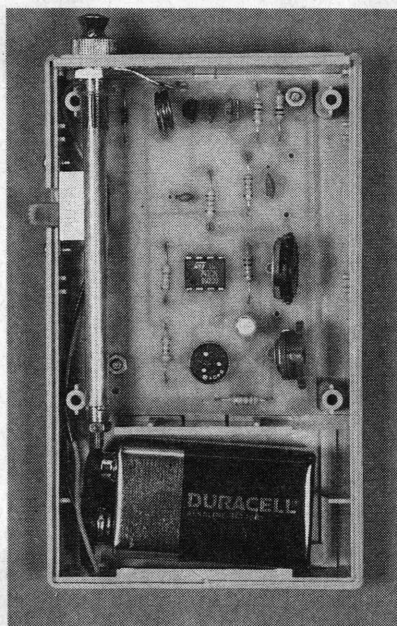


FIG. 10—HERE'S HOW THE BOARD fits in the plastic case. Remember to cut the board between the "CUT A," "CUT B," and "CUT C" points indicated in Fig. 6.

and the R1-C1 junction.

3. Check for approximately 4-volts DC on pin 3 of IC1. This voltage should be variable by turning R11.

4. Check for approximately 2.4-volts DC on the base of Q1.

If any of the voltages in steps 2, 3, or 4 are missing or the values are wrong, you have an open or short circuit.

5. You can check for an oscillation at the emitter of Q1 by using either a frequency counter or oscilloscope. Make sure that the device you use will work at 100 MHz or more. You must also use a $\times 10$ probe to make

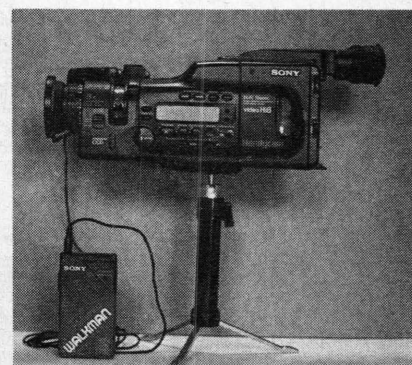


FIG. 11—THE FM RECEIVER CONNECTS to the external microphone input jack on your camcorder.

this test. Any other probe will load down the circuit and kill the oscillation.

If the oscillation is present, but below the FM frequency range of 88 to 108 MHz, you should be able to increase it by turning R11. If you cannot adjust it high enough, then you can compensate with coil L1, which is simply three turns of wire. Separating the windings slightly will lower its inductance and raise the resonant frequency. Adjust L1 as much as necessary until you are able to correct this situation.

If the oscillator does not work, but all the proper voltages are present, then either Q1 or D2 is probably defective.

6. If the oscillator signal is present, but no audio is present, then use an oscilloscope to check for an audio signal at pin 6 of IC1 as you speak into the microphone. Remember, the signal should be riding a DC

USING CONSTANT-CURRENT SOURCES

The constant-current source (CCS) is an overlooked circuit that has a lot of uses in its own right, and makes many other circuits work a lot better. A CCS is a circuit that will sink or source the same current regardless of changes in load resistance or power supply voltage. This article discusses several different types of constant-current sources, including the LM334 CCS diode, circuits based on JFET or bipolar transistors, circuits that use Burr-Brown's REF-series precision reference sources, and configuring three-terminal voltage regulator devices as constant current sources.

The Problem. So why is a constant-current source important? Although there is a very large array of applications for the CCS, the basic problem can be seen in Fig. 1A. The circuit shown there is a voltage divider (or, in instrumentation books, a half-bridge) in which one element (R_2) is a thermistor—a resistor that changes resistance with changes in temperature. The output voltage, $+V_O$, is found from the standard voltage divider equation:

$$V_O = (V_1 \times R_2) / (R_1 + R_2)$$

What that equation tells us is that the output voltage is a function of three factors, not just one. While we would like V_O to change only when R_2 changes (which means that the temperature has changed), variations in source voltage V_1 and the value of R_1 also affects the output voltage. While we can regulate V_1 , and select a high precision, low temperature coefficient resistor for R_1 , there is another way.

That other way is shown in Fig. 1B. In this improved circuit, resistor R_1 is replaced with a CCS. The output current, I_{OUT} , will remain constant even when V_1 and the ambient

Constant-current sources are easy to use, and can make other circuits work better. Here's how.

JOSEPH J. CARR

temperature varies. The output voltage, i.e. the voltage appearing across R_2 , is simply the product $I_{OUT} \times R_2$. Other examples will be considered later when we look at a specific CCS device that you can buy very cheaply.

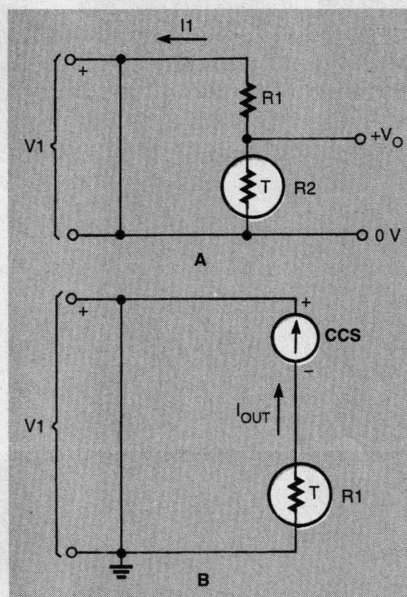


Fig. 1. The circuit in A is a classic half-bridge or voltage-divider circuit in which one element is a thermistor. In the circuit in B, the resistor has been replaced by a CCS.

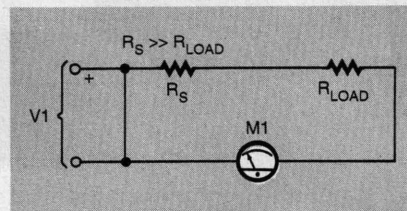


Fig. 2. A crude way to make a CCS: Make the source resistance very, very large compared to the load. Unfortunately, in practice this is usually an ineffective approach.

Constant-Current Source Circuits.

The requirement for a CCS is that the output current does not change even if the ambient temperature, input voltage, and/or the load resistance changes. Figure 2 shows a crude and NOT recommended way of obtaining the CCS action. It does that by making the resistor (R_S) in series with the load (R_{LOAD}) very, very high relative to the load. Consider, for example, the case where $V_1 = 10$ volts, R_S is 1,000,000 ohms, while R_{LOAD} varies from 0 to 200 ohms, with a nominal value of 100 ohms. When the load resistor is 0 ohms, the current flow is $10 \text{ V} / 1,000,000 \text{ ohms} = 0.00001 \text{ A} = 10 \text{ }\mu\text{A}$. When the load resistance rises to 100 ohms, the current is $10 \text{ V} / 1,000,100 = 0.000009999 \text{ A} = 9.999 \text{ }\mu\text{A}$. And when the load resistance rises to 200 ohms, the current is $9.998 \text{ }\mu\text{A}$. Because the source resistance is so large compared with the load resistance, the range of current varies only from 9.998 to 10.00 μA , which is less than 0.02 percent change—constant for practical purposes.

There's only one little hitch to the approach of Fig. 2: currents higher than the microampere level require large voltages. If you need 1 mA, for example, a 1000 volt DC power supply is required for V_1 ! There must be a better way.

Figure 3 shows one better way: use a pair of NPN bipolar transistors cross-connected to provide the constant-current action. That circuit uses a pair of MPS6523 devices, which can be replaced by NTE123AP or

ECG123AP devices. With the values shown, the constant output current (I_O) is approximately:

$$I_O = 0.6/R1$$

Or, with the value shown in Fig. 3 (100 ohms), $I_O = 0.006$ mA. Resistor R1 can be adjusted to provide the exact current required. I used this circuit several times and found that it would output currents from about 500 nA to 10 nA before falling apart and acting like something other than a CCS.

Another "better idea" is shown in

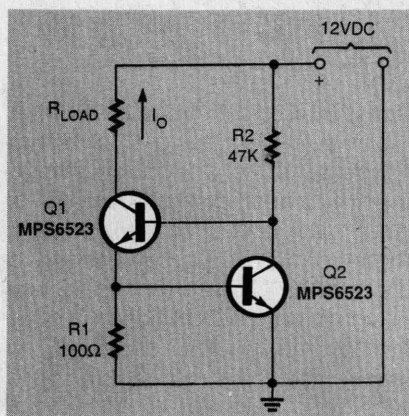


Fig. 3. A better approach is to use two cross-connected NPN bipolar transistors to create a CCS. Standard replacement devices (see text) can be used in place of the transistors shown if desired.

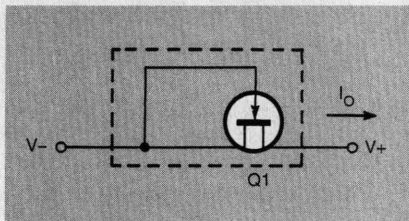


Fig. 4. A JFET can be used as shown to create a current-regulated-diode CCS.

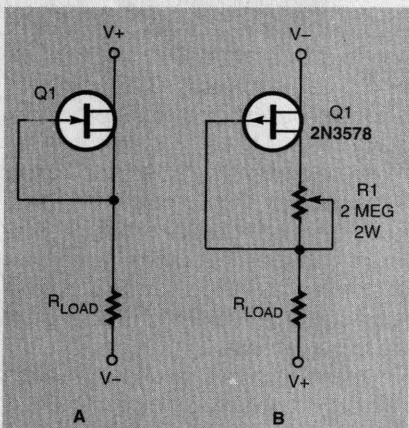


Fig. 5. One way to use the JFET CCS is to place it in series with the load (A). A variable set up is shown in B.

Fig. 4. That circuit uses a diode-connected n-channel junction field-effect transistor (JFET) as a current-regulated diode (CRD) type of CCS. In that approach, the source and gate terminals are tied together to form the (-) electrode of the CCS, and the drain of the JFET forms the (+) electrode of the CCS.

Figure 5 shows two ways that the CRD/CCS is connected into circuits. In Fig. 5A, the CRD is connected in series with the load, in a manner similar to the methods shown earlier. That makes good sense if the current created naturally by the CRD is what you need (or close enough to not matter much). A variable scheme is shown in Fig. 5B. That circuit is adjustable over the range 5 mA to 2 mA when a 2N3578 or its equivalent is used for Q1. Note that the JFET in this case is a p-channel device, so the polarity of the DC power supply, which should be greater than 18 volts in both examples, is reversed from the previous circuit. If you want to make a CCS for use with a positive power supply, experiment with n-channel JFETs with characteristics similar to the 2N3578 device. Note

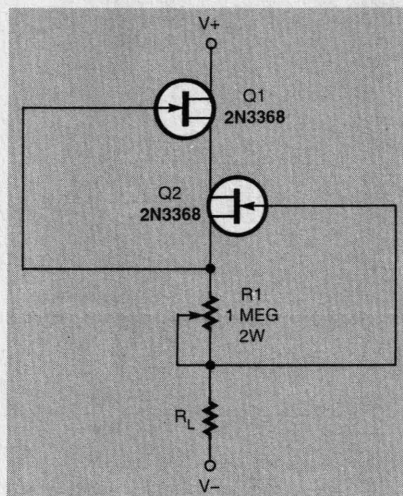


Fig. 6. This adjustable JFET-based CCS uses two transistors and a positive supply.

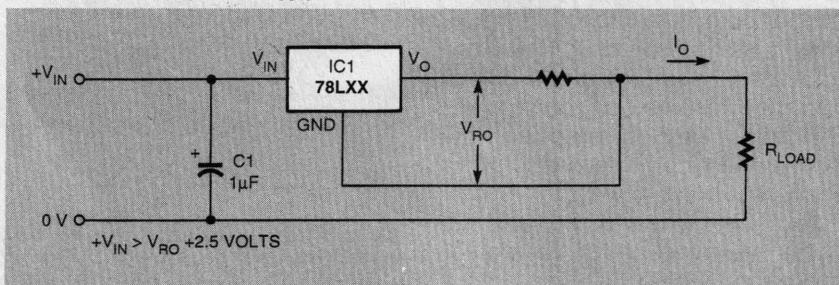


Fig. 7. Ordinary voltage regulators such as the 78Lxx can be used as a CCS when configured as shown here.

that the 2-megohm potentiometer (R1) is a 2-watt type, not the usual 0.5-watt types normally seen.

Another approach to creating a CCS that uses a positive power supply is the circuit shown in Fig. 6. That circuit uses a pair of 2N3368 (or equivalent) n-channel JFET devices cross-connected to form a CRD circuit. With the value of R1 shown, the circuit is adjustable over the range 200 nA to 1 mA. In experimenting with the circuit, I found that it worked better when Q1 and Q2 shared the same thermal environment, which means being very close together on the printed-circuit board. For best stability, one might want to bond the transistors together, or find a dual JFET that has approximately the same characteristics.

IC Current Regulators. In this age of integrated circuits, one would expect that someone would make a device for all purposes, and that's true of the CCS as well. There are several ways to use IC devices to provide constant currents to varying loads. One of the simplest is to use an ordinary 78xx, 78Lxx, 79xx, or 79Lxx three-terminal IC voltage regulator connected as shown in Fig. 7. In that circuit, a 78Lxx (members of the 78Lxx series are 100-mA versions of the 78xx devices) has a resistor (R1) connected between the ground and output terminals, to form a common output terminal to the load. The other end of the load is connected to the input voltage side of IC1 and the 0 V return line on the DC power supply. The output current is approximately $V_{RO}/R1$, where V_{RO} is set by the particular regulator (determined by the "xx" in the type number).

A better approach is to use a member of the REF series of precision voltage sources made by Burr-Brown. The pinouts for two of those

devices are shown in Fig. 8. The REF102 (Fig. 8A) is the replacement for the popular but recently discontinued REF01; it has a 10V output

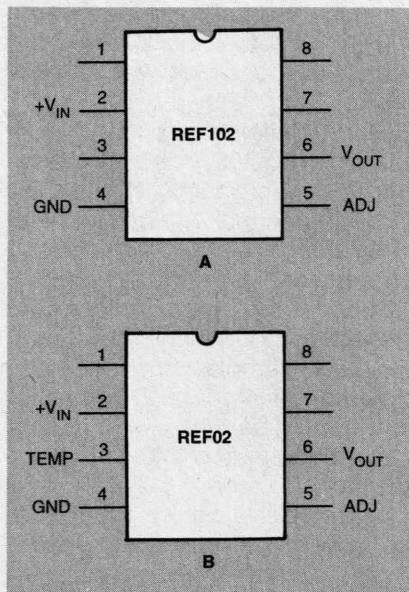


Fig. 8. Packages and pin-outs for the REF102 and REF02 precision voltage-reference ICs from Burr-Brown.

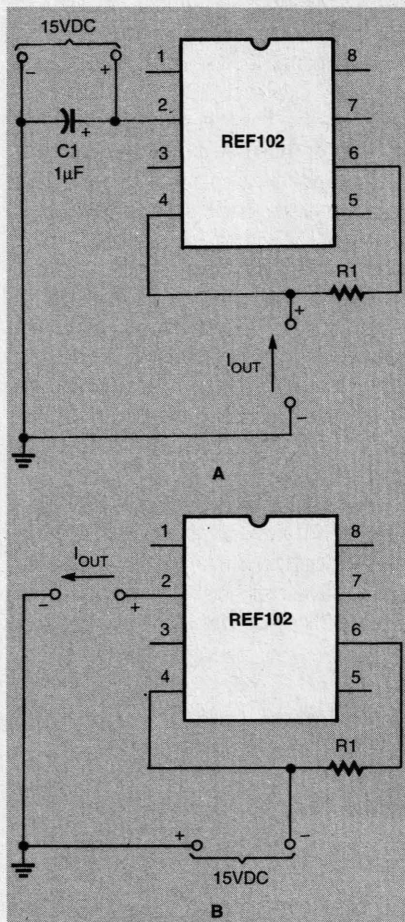


Fig. 9. Here are the circuits for using the REF102 (or REF02) as a current source (A) and sink (B).

voltage (V_{OUT}). The REF02, shown in Fig. 8B has a 5-volt output and adds a TEMP output at pin 3. The voltage appearing at that output is scaled at 2.1 mV/°K, making it usable as a Celsius or Kelvin thermometer.

Figure 9A shows the REF102 connected with a positive power supply, while Fig. 9B shows the negative supply connection. The REF02 could also be used in the same circuit. With either device, the output current is:

$$I_O = (V/R1) + 1\text{mA}$$

Where: I_O is in mA, V is in volts and R1 is in ohms. The values for V are the values of output voltage for the particular device ($V = 10$ for the REF102 and $V = 5$ for REF02).

The LM334. One of the easiest CCS devices to use is the LM334. The device comes in both TO-22 metal and TO-92 plastic packages, although the commonly available form is the TO-92; a pinout of that

package is shown in Fig. 10A. The LM334 operates over a very wide voltage range: 1 V to 40 V, keeping the current relatively stable over the entire range. With one external resistor (Fig. 10B), the LM334 device can produce a stability of 0.02%/V, with a temperature coefficient of 0.33%/°C and a set current range of 1 mA to 10 mA. The initial current accuracy is $\pm 3\%$, but that can be trimmed.

The sense voltage used in the LM334 is proportional to the absolute temperature in degrees Kelvin (°K), i.e. 64 mV/°K, which are the same size as degrees Celsius (°C), but are referenced to absolute zero ($\approx 273.2^\circ\text{C}$) rather than the freezing point of water. That means that an LM334 can be used as a temperature sensor of sorts (more on that later). The initial accuracy is not guaranteed, as in the LM335 temperature-sensor device, but it is good enough for many purposes and can be trimmed with an exter-

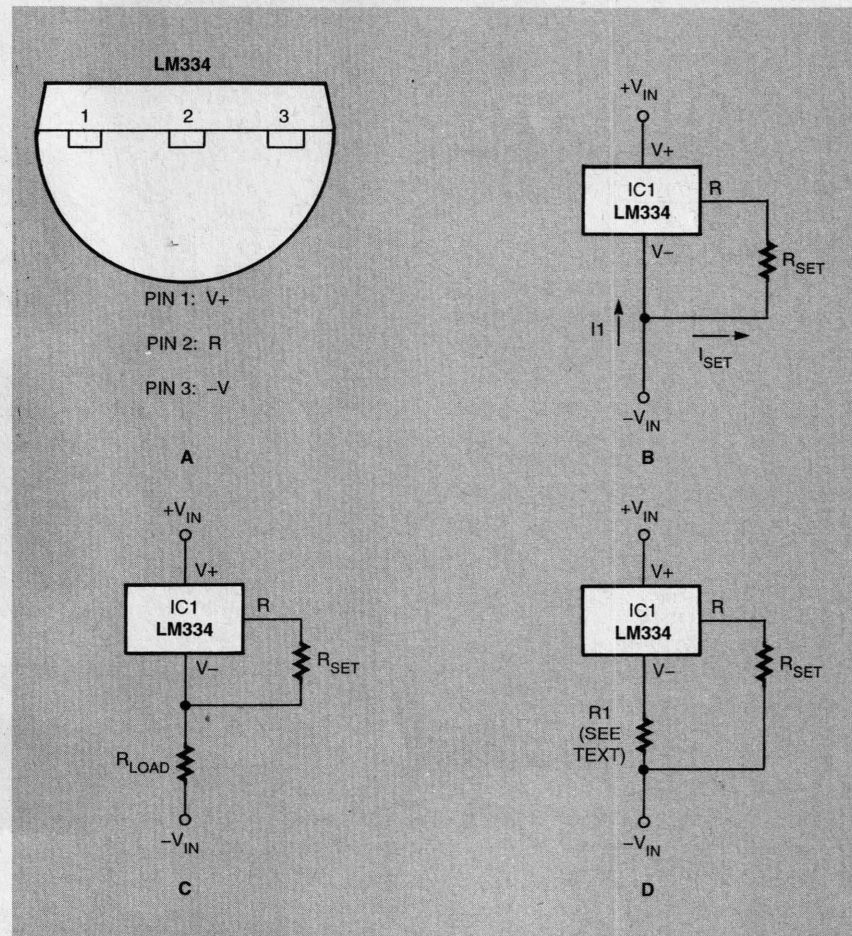


Fig. 10. The easiest to use CCS is the LM334. Its package and pinout are shown in A. A basic circuit is shown in B. The LM334 is shown connected in a circuit with a load resistor in C. An alternate circuit configuration is shown in D.

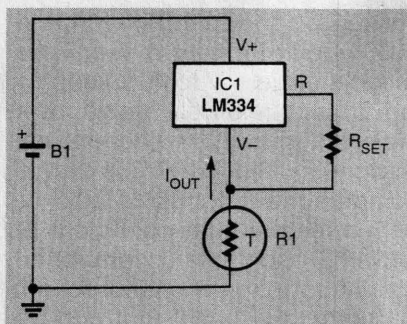


Fig. 11. The performance of the thermistor circuits in Fig. 1 can be improved by using the LM334 as the CCS.

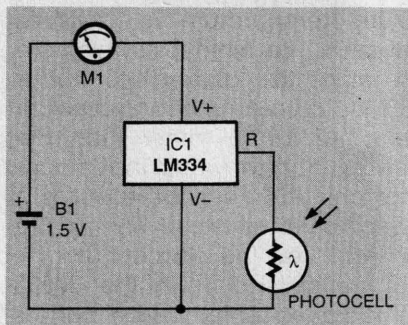


Fig. 12. The LM334 is used here in a light-meter application.

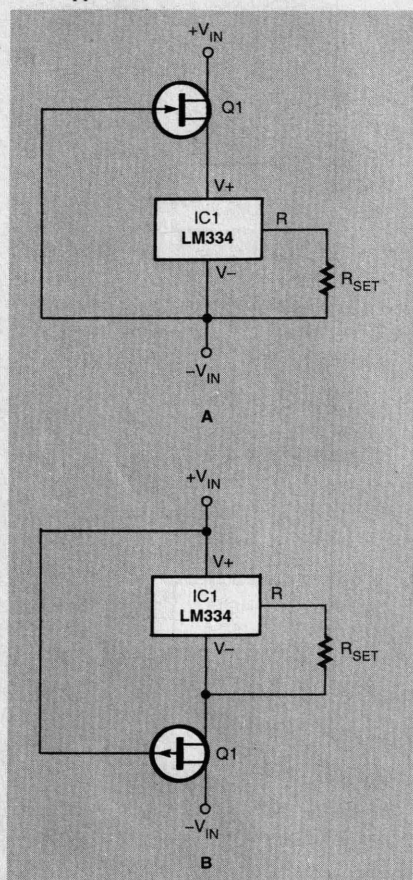


Fig. 13. In these circuits, an n-channel (A) and p-channel (B) JFET is used to reduce the circuit's distributed capacitance and make it look like a much higher source impedance to following circuits.

nal operational amplifier.

The value at the V- terminal current is a fixed ratio with the set current, I_{SET} , which ranges from $I1 = 14 \times I_{SET}$ to $I1 = 16 \times I_{SET}$, depending on the current range and the particular device selected (use a nominal value of 15 for planning purposes). The set current is found from:

$$I_{SET} = 67.7 \text{ mV}/R_{SET}$$

For an R_{SET} of 220 ohms, for example, I_{SET} is $67.7/220 = 0.31 \text{ mA}$, so the output current is $15 \times 0.31 \text{ mA} = 4.65 \text{ mA}$. The value of R_{SET} can be adjusted to sort out differences caused by variation between devices, or the normal variation seen by current

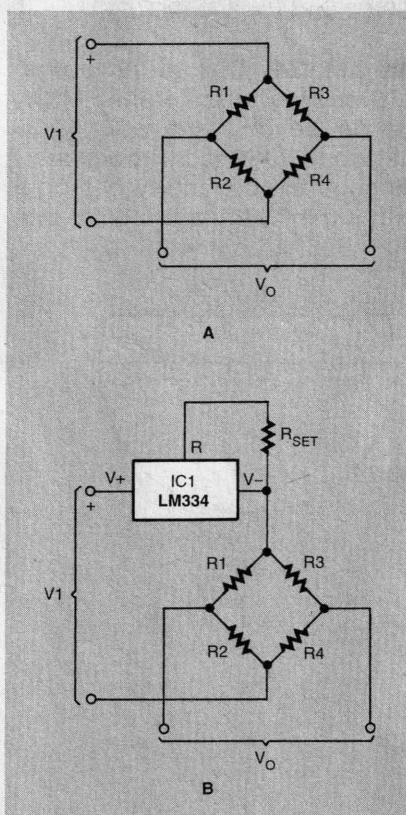


Fig. 14. A standard Wheatstone bridge (A), and one in a CCS configuration.

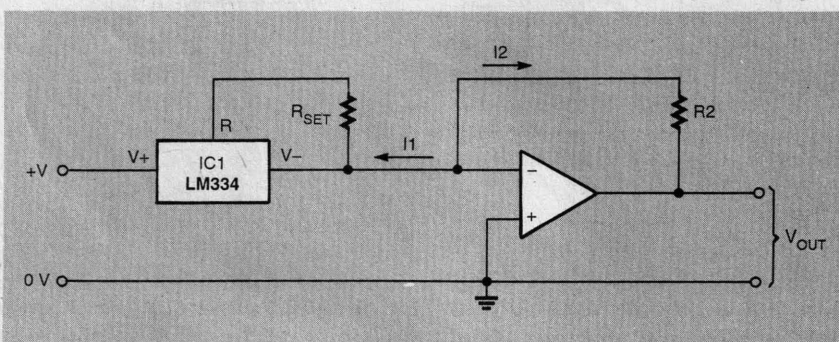


Fig. 15. A CCS can be used with an op-amp to produce output voltage.

range. Once set, it will remain constant so long as R_{SET} doesn't change. That last requirement means that a low temperature-coefficient resistor is needed for R_{SET} . Figure 10C shows the LM334 connected into a circuit with a load resistor.

An alternate circuit configuration is shown in Fig. 10D. In that circuit an extra resistor is added ($R1$) in addition to the normal R_{SET} . To make the circuit work, select R_{SET} to be about ten percent higher than is normally needed for the level of output current required, and then make $R1 = 3 \times R_{SET}$. For example, in the circuit discussed above, R_{SET} was 220 ohms, so here it would be $1.1R_{SET} = (1.1)(220) = 242 \text{ ohms}$ (use a 240-ohm standard value resistor).

Earlier we saw a thermometer circuit in which a thermistor was used as the temperature-sensing element. That circuit can be improved by using a CCS in place of one resistor (as was shown). In Fig. 11, we see a simple circuit in which an LM334 CCS is used to provide the current. For example, if the thermistor resistance is around 10K (a common value) in the middle of the range of interest, a 10 mA constant current would yield a voltage of 100 mV. The problem with that simple circuit is that the LM334 must be kept at a constant temperature. That normally presents little problem if the LM334 is inside an equipment cabinet mounted on a printed circuit board with other circuitry. Enough heat will be generated after the circuit stabilizes to result in a reasonably constant circuit.

Figure 12 shows a circuit for using the LM334 with a photoconductive cell in a light meter circuit. A photoconductive cell has a resistance that is a function of the light shining on

the active surface: increase the light and the resistance drops. If a photo-cell is selected in which the resistance range is compatible with the required R_{SET} range, then the photo-cell can be used in place of R_{SET} .

Figure 13 shows two additional ways of using the LM334. In those circuits, either an n-channel (Fig. 13A) or p-channel (Fig. 13B) JFET device is used in series with the LM334. Because there is a voltage drop across the JFET in each of those circuits, the JFET and operating voltage must be selected to ensure that the voltage drop across the LM334 is at least 1 volt, and not more than 40 volts. With the JFET/LM334 configurations shown, the effect is to reduce the distributed capacitance and to make the device look like a much higher source impedance to following circuits.

Figure 14 shows two ways to excite a Wheatstone bridge. In Fig. 14A, the bridge is excited by a voltage source, V1. That works well in most cases. An alternative method is to use the LM334 in series with the voltage source (Fig. 14B). The available current from the LM334 splits between the two paths through the bridge circuit. It is common practice to make all resistors equal, and then vary one or more resistors in response to an external stimulus. For example, R1, R2, and R3 could be fixed resistors, while R4 could be a thermistor, photoconductive cell, or piezoresistive strain gage. The Wheatstone bridge is superior to the voltage divider method because the output voltage drops to zero when the stimulus is zero, assuming that the resting resistance of R4 causes the circuit to obey the ratio $R1/R2 = R3/R4$. The voltage drop across each resistor is a function of the constant current flowing in the circuit.

A final application for the LM334 (or any other CCS, for that matter) is shown in Fig. 15. In that circuit, the CCS is connected into the inverting input of an operational amplifier. The output voltage (V_{OUT}) is found from the product of $I1$ and $R2$, or $I1 \times R2$. The output voltage will be constant so long as $R2$ is constant, but if $R2$ is a resistance-based sensor (e.g. thermistor, photoconductive cell, etc), then the output voltage is set by the

change of resistance, which is in turn set by the applied stimulus.

Well, that about wraps things up. As you can see from this article, the constant-current source solves a lot of circuit problems, and is relatively easy to use. It is well worth learning about as it is the basis for a large collection of circuits used in instrumentation and other fields. Ω

DAYTIME RUNNING LIGHTS

(continued from page 52)

frame location to the battery's negative terminal; the resistance should be less than 2 ohms. Securely mount the DRL Adapter to the car.

Find a source of 12-volt power that can provide up to 15 amps of current and is only on when the ignition switch is in the "on" position. You can test the wire by carefully inserting a needle into the wire and measuring the voltage with a voltmeter while a helper switches the ignition on and off. Don't forget to check that the wire does not carry any current when the ignition switch is in the "accessory" position.

Use a wire splice to connect the 12-volt wire from the DRL Adapter to the switched 12-volt wire that you just found. Next, find a headlight connector that is easy to access and locate the wire that will power the low beams. That might not be as easy as it sounds. On some cars, high-beam/low-beam switching is done from the grounded side of the circuit. If that is your situation, connect the DRL Adapter to the common lead of the headlights. Splice the headlight wire from the DRL Adapter into that headlight wire.

Before routing and tying off the cables, test the DRL Adapter by turning on the ignition switch and making sure that the headlights do not turn on. Start the engine, and the headlights should come on at somewhat less than full brightness. That is easily tested by turning the headlights on and off with the motor running. When you turn the headlights on from the dashboard, they should come up to full intensity and return to 80% brightness when the headlight switch is turned off.

With everything tested and veri-

PARTS LIST FOR THE DAYTIME-RUNNING LIGHTS ADAPTER

SEMICONDUCTORS

IC1—LM555 bi-polar timer, integrated circuit
D1—1N759A Zener diode, 12-volt
D2—1N4001 silicon diode
Q1—2N3906, silicon transistor, PNP
Q2—2N6053, silicon transistor, PNP Darlington
Q3—2N3904, silicon transistor, NPN

RESISTORS

(All resistors are 1/4-watt, 5% units.)
R1—150-ohm
R2—330-ohm
R3, R6—1000-ohm
R4—4700-ohm
R5, R7—10,000-ohm

CAPACITORS

C1—0.1- μ F, ceramic-disc
C2—22- μ F, 25-WVDC, electrolytic
C3—0.001- μ F, ceramic-disc
C4—0.05- μ F, ceramic-disc

ADDITIONAL PARTS AND MATERIALS

J1—2-pin connector
Case, wire, insulated standoffs, hardware, etc.

Note: The following items are available from: Harrison R&D, 9802 Sagequeen, Houston, TX 77089; Tel: 281-485-7107; e-mail: dharrison@ghgcorp.com; Kit of all parts including PC board, \$30; PC board only, \$7. Please include \$3.50 for shipping and handling within US and Canada. Texas residents must add 6 1/4% sales tax.

fied to be working properly, route the wires and tie them so that they will not rub against any sharp edges or touch any hot parts on the engine. With daytime-running lights, everyone will be safer. Ω

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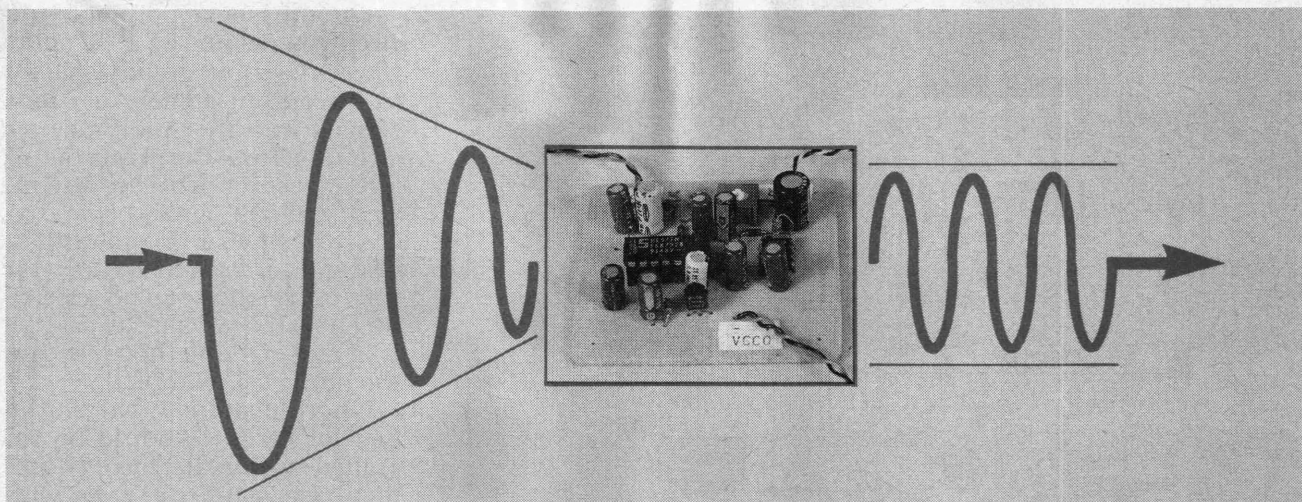
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AUDIO LEVEL CONTROLLER

Build this programmable audio-level controller to tame your receiver's input and keep it in your ear's comfort zone.

STEVE SZABO, N1AYO

HAVE YOU EVER BEEN STARTLED when the audio from your radio receiver went out of control and nearly blasted you out of your chair? Maybe you just turned up the volume in an effort to hear a distant station or a soft voice, and within seconds an unwanted burst of sound overwhelmed you. If your answer is yes, you need an automatic audio level controller.

Not all transceivers and scanners—even the most expensive high-end products—have automatic volume control circuits. Now you can build the Audio Leveler module to keep your receiver's audio volume constant, regardless of incoming signal strength. It will work in your ham receiver, scanner, marine or other mobile transceiver, television set, or stereo system.

The Audio Leveler is independent of the host receiver's volume control. You set its volume to a comfortable level, and the Audio Leveler locks in on that setting. It amplifies desired low-level audio signals while discriminating against background noise and attenuating strong random signals.

Circuit description

The heart of the Audio Leveler is a Signetics NE577, a programmable, low-power integrated circuit called a compandor. The NE577, shown as a simplified block diagram in Fig. 1, is packaged in a 14-pin DIP. The first question you might well ask is: What is a compandor, and what does it have to do with audio leveling? The answer is that it's a circuit capable of compressing and expanding an input signal to remove noise in a communications channel, and one of its sections can be organized to control audio input signal level. The term is derived from a contraction of the two words *compressor* and *expander*.

The compandor was developed as a discrete component circuit for telecommunications applications, primarily to reduce unwanted noise. The input signal is fed to a compressor stage which rectifies and conditions it so that the input signal level always remains above the noise level. The conditioned signal is then fed to the expander stage which restores it to its ini-

tial volume level—any noise is expanded below the audible level.

The NE577 has both a compressor and an expander stage, but the Audio Leveler uses only the compressor stage which is configured as a programmable automatic level control (ALC). The ALC accepts a range of input signals, and produces a constant AC output level. The host equipment can have a volume control, but with the Audio Leveler in your receiver, you will only have to set the receiver's volume control once.

Figure 2 is the schematic for the programmable Audio Leveler. Only the compressor section pins on IC1, the NE577, are used; pins 1 to 3 in the expander section are not used. The ALC function is configured with the rectifier at $RECT_{in}$ pin 10 and $GAINCELL_{in}$ at pin 9 forming a closed loop around the internal op amp. Because the AC output level of the ALC can be programmed, you can choose a resistor value for a desired output level.

The audio signal is fed simultaneously to both pins 10 and

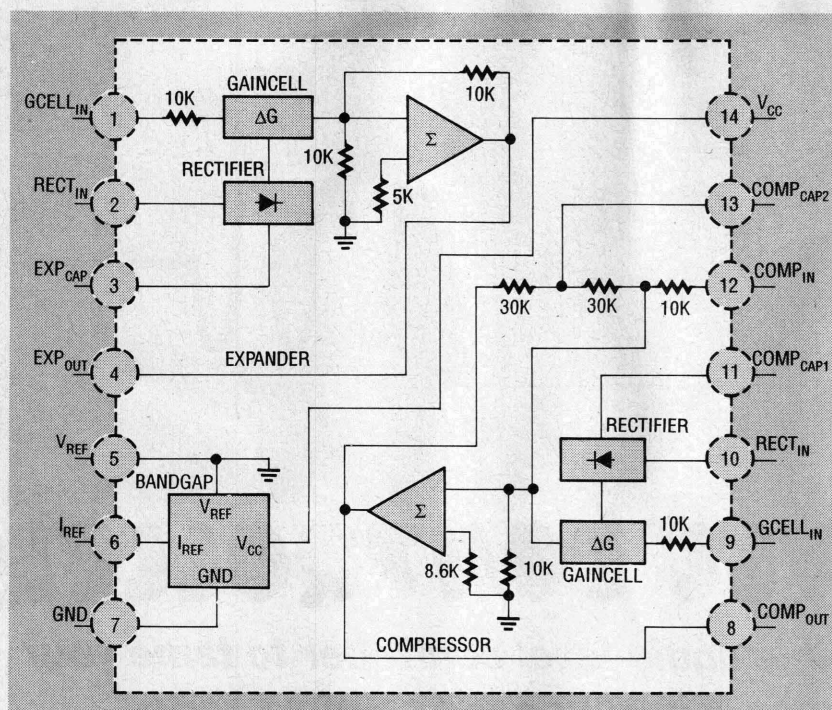


FIG. 1—SCHEMATIC FOR THE AUDIO LEVELER. The audio output stage is not needed if you can use the host receiver's audio amplifier stage.

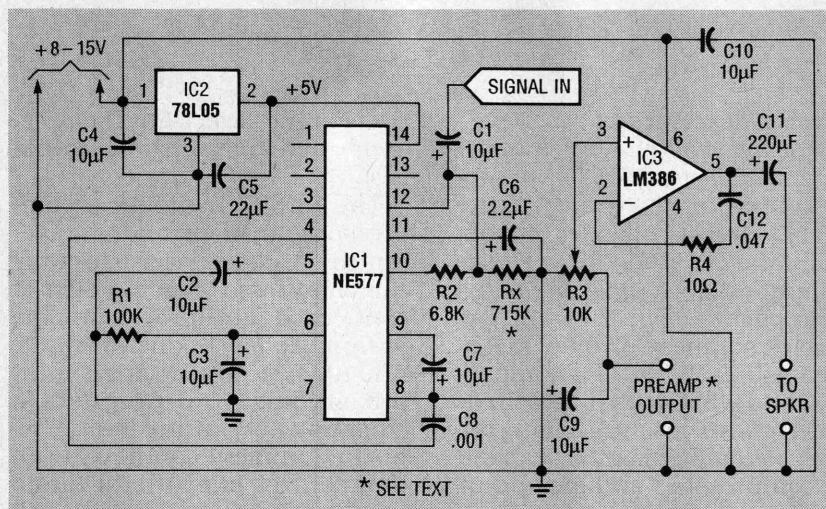


FIG. 2.—BLOCK DIAGRAM OF THE NES77 LOW-POWER COMPANDOR. This IC compresses the input signal to eliminate noise and control range, and then expands it again for normal listening

12. At pin 10, resistor R2, with a value of 6.8 K, is shunted by programmable resistor R_X , which limits maximum gain to prevent amplification of background noise. If resistor R2 is 6.8 K, R_X can be selected so that an input signal below 10 millivolts will not be amplified with a gain greater than 10. The circuit's output remains at a constant 100-millivolt (rms) level for the range of input voltages shown in Table 1.

The value of resistor R_X can

vary from 64.3 K to 715 K, depending on input signal conditions. A value of 715 K for R_X was obtained in the prototype by connecting a 680 K resistor in series with a 33 K resistor.

If the input is 10 millivolts (rms) and R_X is 715 K, the maximum circuit gain is limited to 10 for an output of approximately 100 millivolts. Increasing the value of resistor R2 will increase circuit gain and output level, but reducing R_X will reduce the threshold level.

The crossover point of this circuit is defined as that point where the input signal is equal to the output signal. All input signals into the Audio Leveler circuit above the unity gain-level crossover point are attenuated, while all signals below that crossover point are amplified. The optimum threshold level for your application can be selected by choosing a value of R_X , by trial and error methods, within the limits set by Table 1.

Where practical, the Audio Leveler circuit should be installed within its host equipment enclosure in series with its volume control. Your receiver or scanner might not have enough space within its enclosure to accommodate the Audio Leveler, but do not alter the layout of the circuit board to fit a confined space unless you have enough experience to solve any interference, insulation, or thermal problems that might arise.

TABLE 1
DYNAMIC RANGE WITH
DIFFERENT R_X RESISTOR VALUES

For a 100-millivolt rms output for different values of gain:

- (1) Set input voltage, and
- (2) Set R_X as follows:

Input (volts,rms)	R_X (K ohms)	Gain (approx.)
1.0	6.43	1
0.050	136.0	2
0.100	715.0	10

Circuit construction

Conventional electronic circuit construction practice should be followed in building the Audio Leveler. The circuit can be built on standard perforated circuit board with a 0.1-inch grid, but a circuit board is strongly recommended. A foil pattern is provided in this article if you want to make the board. Alternatively, it can be purchased as a separate item from the source given in the Parts List. Regardless of your choice, be sure to drill holes in the corners of the board at the right locations for fastening it

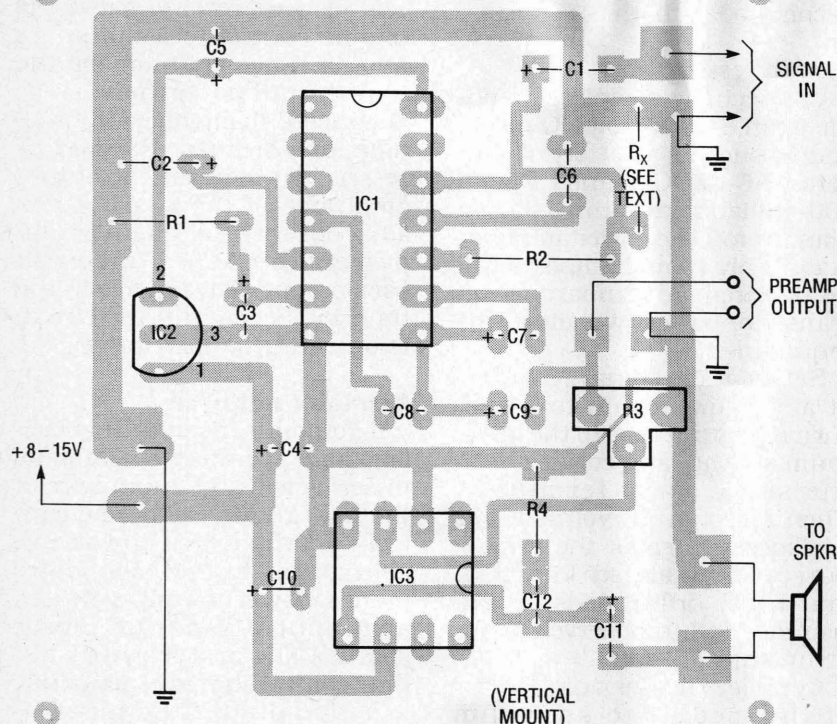


FIG. 3—PARTS PLACEMENT diagram for the Audio Leveler. If the module is bundled with the host receiver, the output stage might not be needed.

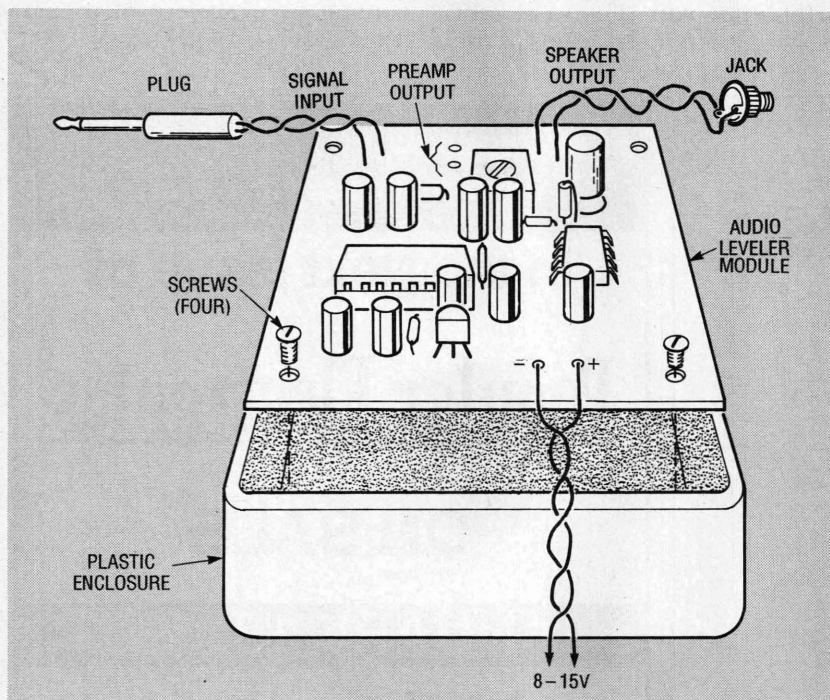


FIG. 4—MODULE ASSEMBLY AND TEST diagram. The stand-alone module can be placed in a separate plastic enclosure with a battery for power.

within your intended enclosure (host cabinet or separate plastic box) before starting assembly work.

Refer to the parts placement diagram, Fig. 3. Insert all resistors and capacitors in their proper places: (Two resistors in

series might be required to obtain the desired value of R_x .) Then insert a recommended socket for IC1. Insert the ends of all input, output and power cable or twisted wires at the terminals shown in Fig. 4.

Carefully check the module

PARTS LIST

All resistors are 1/4-watt, 5%

- R1—100,000 ohms
- R2—6800 ohms
- R3—10,000 ohm, potentiometer, PC board mount
- R4—10 ohms
- R_x —715,000 ohms, 680,000 ohms in series with 33,000 ohms

Semiconductors

- IC1—NE 577 unity-gain, programmable low-power compandor (Signetics) or equivalent
- IC2—LM78L05ACZ 5-volt voltage regulator (National Semiconductor) or equivalent
- IC3—LM386 audio power amplifier (National Semiconductor) or equivalent

Capacitors

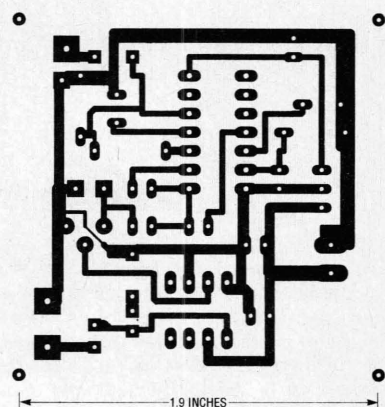
- C1-C4, C7, C9, C10—10 μ F, 16 volts, electrolytic
- C5—22 μ F, 10 volts, electrolytic
- C6—2.2 μ F, 10 volts electrolytic
- C8—0.001 μ F
- C11—220 μ F, 16 volts, electrolytic
- C12—0.047 μ F,

Miscellaneous: circuit board, experimenter's plastic enclosure (optional), power source (see text), twisted-wire pairs or audio cable, solder, jacks and plugs as required.

Note: The following parts are available from C & S Electronics, P.O. Box 2142, Norwalk, CT 06852-2142, phone or fax: (203) 866-3208

- Formed and drilled PC board—\$12.95
- Complete kit of parts excluding power supply and cabinet—\$24.95.
- An assembled and tested module (Model ALC225C)—\$32.95.

Please send check or money order only. Connecticut residents add 6% tax. Add \$3.00 for postage and handling.



FOIL PATTERN for the audio leveler project

for mistakes and inadvertent solder shorts, and make all corrections before inserting and soldering IC2 and IC3. The Au-

dio Leveler is a low-level audio circuit, so trim all leads as short as possible to minimize noise interference. Then insert IC1 in its socket.

The twisted-wire pair from the prototype module's SPEAKER OUTPUT terminals was terminated with a ¼-inch open-frame jack, and the pair from the SIGNAL INPUT terminals was terminated with a ¼-inch plug. The 8 to 15 volts DC are supplied over a twisted pair. Select jacks and plugs that interface directly with those on your receiver or scanner.

If you plan to install the module within a receiver enclosure and do not need the output audio stage, omit the LM386 audio power amplifier, IC3, and related components R4, C11 and C12, and use the PREAMP OUTPUT terminals for the circuit output.

If you plan to use the Audio Leveler as a stand-alone accessory, put it in a separate plastic experimenter's enclosure as shown in Fig. 4. If you want to power the module with a bat-

tery, select an enclosure that will accommodate both.

Circuit testing

Connect the completed module to an 8- to 15-volt DC power source such as a battery or universal AC-to-DC adapter with a 100-milliamperere rating. (If you plan to omit the power amplifier stage, only 10 milliamperes will be required. A standard 9-volt transistor battery will meet this requirement.)

Set an audio signal generator at any frequency up to 1 kHz, turn its output level to the minimum setting, and connect it to the SIGNAL INPUT terminals. Then place an AC voltmeter or oscilloscope across the PREAMP OUTPUT terminals (see Figs. 2, 3, and 4, two drilled pads next to trimmer R3) to observe the instrument's output.

Connect the SPEAKER OUTPUT of the module to an 8-ohm speaker. Set trimmer R3 to an audio output level that is comfortable for you. Then connect the SIGNAL INPUT terminals of the

module to the external speaker jack of your receiver, turn the volume control to a minimum level, and slowly increase the setting until no further change is noticed. To change audio volume, only trimmer R3 need be set. As you increase the output amplitude of the signal generator, observe the circuit's output on the AC voltmeter or oscilloscope. Any change in volume will be slight if the circuit is operating satisfactorily.

Amplifier not used

If you omit the output amplifier stage, connect the module in series with the input wire of the host receiver's volume control—not the wiper. (If you can gain access to the host equipment's power supply, you can also omit a separate power source.) The host receiver's volume control should remain fully functional after the installation. The volume level of the host receiver will stay nearly constant, regardless of incoming signal strength.

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STORM-WARNING LIGHTNING MONITOR

Although discussions of El Niño are no longer front-page headlines, the unpredictable weather—including violent thunderstorms—has always been here and likely always will. For example, an estimated two-billion dollars worth of damage is caused every year by lightning. Some electronics manufacturers are reluctant to make good on warranties if the cause of the failure was due to a lightning strike. With more surge-sensitive equipment coming into everyday use, our world has become more susceptible to “technological shutdown” from lightning hits.

In addition to property damage, lightning strikes are dangerous. Last year, over 200 people were struck and killed by lightning. Your odds of being hit by lightning are said to be better than winning the lottery!

Look around your home. If there is a tall antenna mast or tree on or near your property, the statistical chances of a lightning hit are greatly increased. Although tall, grounded objects around you or your home will tend to protect you from direct hits, the damage done by a nearby lightning strike is not something that should be taken lightly. Even after it gets into the ground, lightning will travel long distances before it dissipates—following buried pipes and cables that act as conductors; the electrical charge can resurface anywhere. For example, although a ham-radio mast might be well grounded, it is always a good idea to disconnect any equipment from the antenna when a storm is approaching.

A direct or nearby strike can take out much more than just the radio. One example that the author is familiar with occurred in an office that decided to do their own local-area-network wiring. They decided to save installation costs by having their in-house electrician run the cables. Several of the cables were



Plug this simple circuit into an AM radio, and you'll know well in advance of any approaching thunderstorms.

hung from the overhead sprinkler pipes. Soon, certain computers began to break down on a regular basis. The common factor between them was that they were all connected to cables that were strung up against hundreds of feet of “grounded” overhead sprinkler pipe. When it was suggested that lightning was the culprit, the electrician argued that grounded pipes should not be above ground potential, even when struck by lightning.

It was eventually verified that the EMF pulses from nearby lightning strikes were being induced into the LAN wire by the sprinkler system. Although the cause of the malfunctions was eventually found, it was at a high cost in both lost hardware and time—all for the desire to save a few dollars!

The Nature of Lightning. Lightning is an electrical discharge that rebalances the differences between pos-

KENTON CHUN

itive and negative charges within a cloud, between two clouds, or between a cloud and the ground. Cloud-to-ground strikes generally cause the greatest amount of property damage and injury to life. The negative charge at the cloud's base is attracted to the positive charge at the earth's surface, which causes the initial contact. An actual lightning strike consists of two separate events—a strike leader and a return stroke. When the electrical potential is great enough, a small strike leader will leap from the cloud to ground, ionizing the air as it goes. The return stroke flows back up the ionization path into the cloud creating the larger flash and discharge that we associate with lightning.

Due to the large voltages involved, a lightning strike creates a broadband EMF pulse of tremendous magnitude. Those pulses can easily be detected a hundred miles away. When a storm approaches,

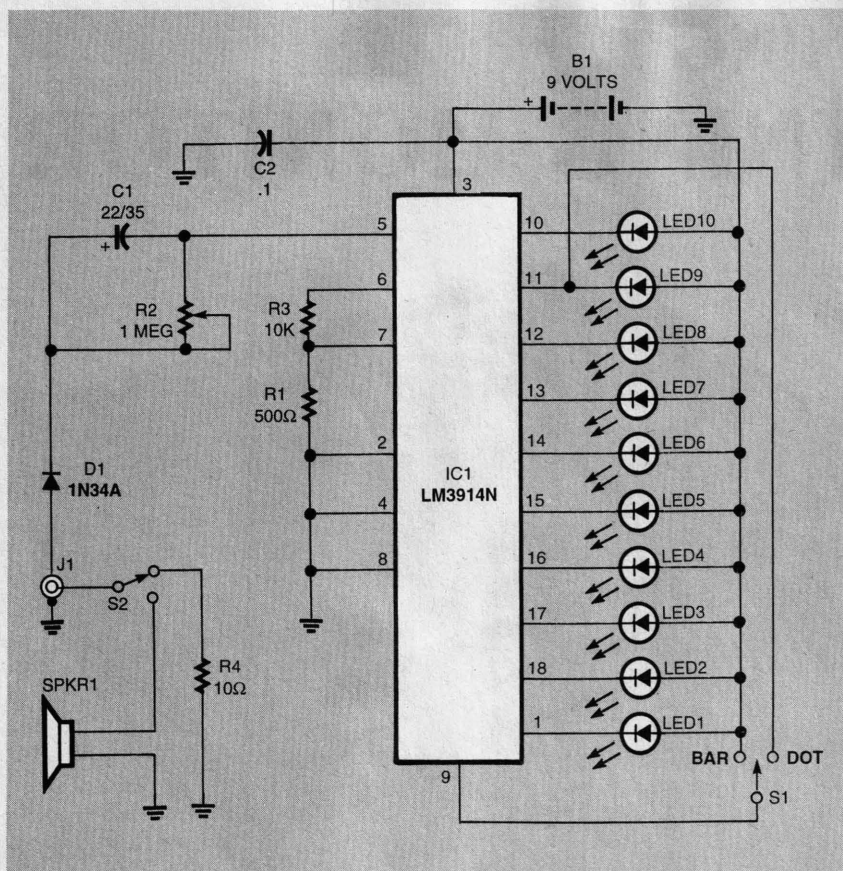


Fig. 1. Built around an LM3914 LED display driver, the Lightning Monitor listens to an AM radio. When a pulse from a lightning strike is detected, the LEDs light up. The amount of time that they remain lit is adjustable so that you have time to notice the warning.

the frequency and field intensity of the EMF pulses will increase. We can take advantage of that phenomenon to "detect" the presence of an approaching thunderstorm.

By nature, a bolt of lightning is an excellent amplitude-modulated (AM) signal. Many of us have heard lightning strikes on an AM radio. Because of the broadband nature of the pulse, the crackle of a lightning strike can be heard anywhere on the AM dial. Although you can simply leave the radio on all of the time, it is impossible to pay attention to it all of the time. Moreover, our sense of hearing is not very acute—our brains have learned to "tune out" noise. To overcome that problem, we need some way to draw our attention to the possibility of an approaching storm. That task is nicely handled by the Lightning Monitor that we will now describe.

How It Works. The Lightning Monitor is a simple circuit that listens to an AM radio signal continuously. The circuit diagram is shown in Fig. 1. The

heart of the circuit is IC1, an LM3914N display driver. That chip is designed to show a voltage reading on a set of light-emitting diodes. Within IC1 are a voltage divider and ten comparators that turn on in sequence as the input voltage rises.

The audio output from an AM radio is applied to J1. If you want to hear the radio signal, S2 switches the audio signal between the Lightning Monitor circuit and SPKR1. When the speaker is not in use, its load on the radio is simulated by R4.

The audio signal is changed into a half-wave DC signal by D1. The semi-rectified signal is applied to C1, charging it up. During the non-charging portion of the half-wave signal, C1 is discharged by R2. Since R2 is adjustable, the discharge rate of C1 can be set to match the audio characteristics of the particular radio being used. The rate of discharge can be varied between almost instantaneously to several seconds.

Normally, the audio signal will hold the voltage on C1 steady with

some fluctuation as the audio gets louder and softer. When a pulse from a lightning strike surges through the radio, C1 charges much faster than it discharges.

The voltage on C1 is sensed by IC1 and the current level displayed on the LEDs. Since IC1 can display a voltage level as a "bar" of LEDs or as a single moving dot, S1 is used to select between the two display modes.

The circuit is powered by B1, a 9-volt battery, although any source of well-filtered DC power between 5 volts and 18 volts will work well. In fact, if the circuit is to be left on all the time, you'll almost certainly want to use an external power source. That's because should B1's voltage drop below IC1's minimum supply requirement, the comparators within that IC, and hence the unit, will not be reliable.

Building the Lightning Monitor. The Lightning Monitor circuit is simple enough to be built on a piece of perfboard using standard construction techniques. The size of the LEDs and their colors depend on your own personal preference. If you want to arrange the display similar to other color meter displays, use green LEDs for the first six units, followed by two yellow LEDs and two red LEDs at the high end. All of the connections should be kept as short as possible.

The Lightning Monitor can be housed in any suitable case. If you are exceptionally neat in your board construction, you can use a clear plastic box similar to the author's prototype shown in Fig. 2. If your radio has enough room inside its case, you can even mount the Lightning Monitor inside that case and use the radio's power supply to power the circuit.

Calibrating and Using the Lightning Monitor. Tune a radio to the quietest spot that can be found on the AM band on a day with clear weather. It will probably be towards the higher end of the band—the lower end of the AM broadcast band tends to be more susceptible to terrestrially-generated EMI and RF interference. You might want to do that over a few

(Continued on page 47)

to lightning, you should think about lightning protection. A direct lightning hit on the Digital Wind Vane will probably destroy everything from the encoder to the cable to the display unit. Be aware that even nearby lightning strikes could also damage the unit's electronics—particularly when using long cable runs. Pull-down resistor R2 gives some protection from static build-up. As an added precaution, connect the power supply's negative lead to a good earth ground.

Even if you are just adding an encoder to a wind vane that has been up for years, you now have a lead-in cable that will conduct lightning directly into your house. You should make sure that the mast is properly grounded or install a grounded lightning rod that is higher than the wind vane and mast. Another technique is to bend the input cable around some corners before it enters the house. Lightning usually will not take a corner as long as it has a straight ground path to follow. Furthermore, you can also use discharge tubes on each cable wire. Ω

LIGHTNING MONITOR

(continued from page 40)

days in order to get a good idea of how quiet the target frequency really is. AM broadcasts and interference often live a strange nomadic existence on the AM band.

Power up the Lightning Monitor and connect it to the speaker or headphone jack of the radio. Ideally, the radio should cut the speaker out when the Lightning Monitor is connected.

Set S2 to the "audio" position so that the radio can be heard through SPKR1. If a thunderstorm is in the vicinity, you should be hearing the static crashes from lightning strikes. If reception is good, you might even be able to hear them from thunderstorms that are many miles away. If a thunderstorm is not in the area, simulate one by using a hairdryer to create static or tune the radio to a strong station.

Set the radio's volume control so that the monitor gives a full-scale reading on the loudest of strikes.

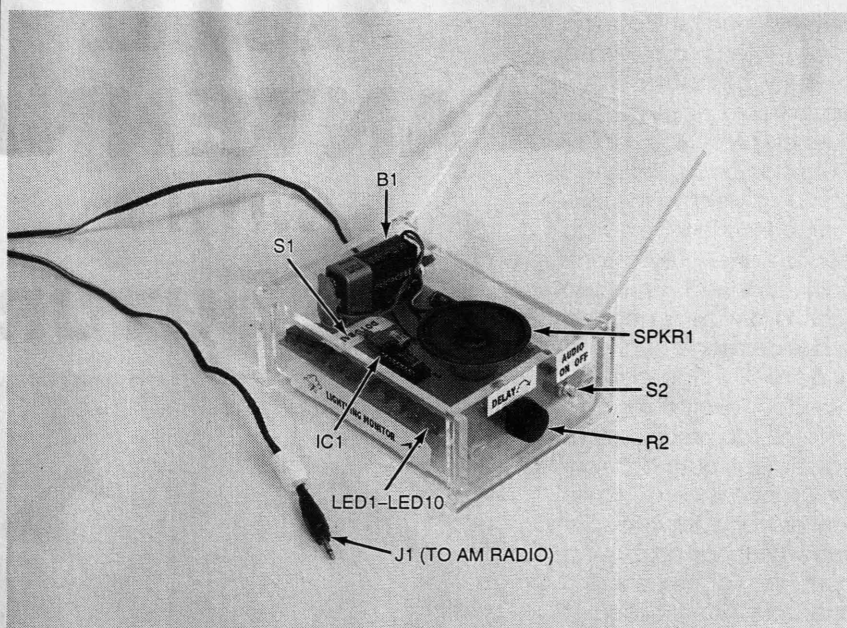


Fig. 2. The Lightning Monitor is simple enough to build on a perfboard. Here, the author's prototype has been mounted in a clear plastic case, making the Lightning Monitor both a decorative and a practical project. The battery can be replaced by a wall-mounted transformer for continuous use.

PARTS LIST FOR THE LIGHTNING MONITOR

SEMICONDUCTORS

IC1—LM3914N LED display driver, integrated circuit
D1—1N34A germanium diode
LED1—LED10—Light-emitting diodes (see text)

RESISTORS

(All resistors are 1/4-watt, 5% units unless otherwise noted.)
R1—500-ohm
R2—1-megohm, potentiometer
R3—10,000-ohm
R4—10-ohm

CAPACITORS

C1—22- μ F, 35-WVDC, electrolytic
C2—0.1- μ F, Mylar

ADDITIONAL PARTS AND MATERIALS

B1—9-volt battery (see text)
S1, S2—Single-pole, double-throw switch
SPKR1—2-inch speaker
Case, knob, AM radio, hardware, etc.

Finally, set R2 so that the readings "fade out" inside a time interval that is reasonable for you to notice the event. A few seconds is usually sufficient. For continuous use, flip S2 to the other position and you will have peace and quiet again. If you begin to see readings on the LED display, flip S2 back to the "audio" position and listen to confirm that a storm is approaching—and not a local unli-

cenced "pirate" AM station.

If you live in an area where the standard AM broadcast band is too noisy, you might want to use a short-wave, marine, or CB radio. At frequencies over 15 MHz, terrestrial noise is less noticeable.

If you build the Lightning Monitor into a small box, you can also use it on days of clear weather as a portable "storage voltmeter" for measuring slow-changing low-voltage events such as temperature changes or light levels with a suitable sensor. Even if you don't happen to live in the "lightning capital of the world," it is always best to be prepared! Ω

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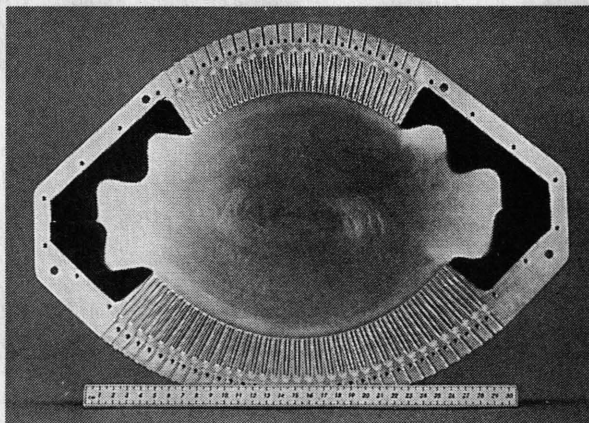
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The first Rotman lens to operate at millimeter-wave frequencies (i.e. frequencies as high as 37 GHz) has been designed and built, yielding the prototype for a low-cost, high performance, electronically scanned antenna. Such a device is needed in a wide range of civilian and military applications, including aircraft landing systems, communications equipment, auto collision-avoidance systems, missile seekers, and tank radars. Because it has no moving parts, no phase shifters, and can be encased in plastic, researchers at the Georgia Institute of Technology, where the device was designed and built, believe they have an inexpensive, rugged, reliable, and compact alternative to current millimeter-wave (MMW) antenna technologies.

"MMW components are compact and well suited for integration into missile-seeker heads, smart munitions, automobile collision-avoidance systems, and synthetic-vision systems," said Georgia Tech's Ekkehart (Otto) Rausch, senior research scientist. "In these applications, low-cost, rapid, inertialess scanning of the antenna is desirable. Most MMW antennas that operate at frequencies equal to or greater than 35 GHz use either a mechanical scanning approach or phase shifters for electronic steering. Phase shifters that operate at MMW frequencies are costly and introduce considerable RF loss. Mechanically steered antennas are relatively low in cost but are slow in response, sensitive to shock and vibration, and have moving parts that are subject to wear and failure. Thus, low-cost, high-reliability, and electronic scanning are generally incompatible unless the design is based on a MMW Rotman lens."

By avoiding those limitations, the new Rotman lens could open new applications for MMW radar.

A SMALL ANTENNA WITH A BIG FUTURE



New technology has yielded a low-cost, high-reliability antenna that can operate at frequencies to 37 GHz.

DOUGLAS PAGE

How it Works. Rotman-lens devices get their names from their ability to focus microwave or millimeter-wave energy coming from a particular direction by passing this electromagnetic energy through a pair of lens-shaped parallel plates.

"This Rotman lens consists of a parallel plate region with waveguide ports distributed around the periphery of the plate," Rausch explains. "Beam-forming or focal ports are located on one side of the plates. Those ports are fed by a switch array. The array ports are on the opposite side, each connected to an antenna element. Energy, when input into a specific focal port, will emerge from the antenna elements and produce a beam along a particular direction. Switching the input from focal port to focal port will steer the beam electronically in one dimension. The concept may be extended to two dimensions by modifying the Rotman lens equa-

tions and generating two-dimensional surfaces for the focal and array contours."

Since the objective of the research was proof-of-concept, the design was restricted to one-dimensional scanning. Currently, the lens is constructed of aluminum. The potential exists, however, to build both the lens and the switch array using plastic. Production costs can be kept low by hot-pressing the antenna in plastic, which could then be coated with a conductor, such as gold. The antenna feed horns and switch arrays would be assembled the same way.

In addition to low cost, durability and compact size, another feature of the Rotman lens is the low throughput loss and sidelobe emissions. Sidelobe power in the prototype can be suppressed by a factor of one-thousand below the energy of the main beam. The lens itself loses less than 2 dB power.

"The antenna we are designing for Phase II of this project will have a gain in excess of 30 dB," said Rausch. "This antenna gain will be achieved with an array of 32 horn-antenna elements having a beamwidth of less than four degrees in azimuth and less than eight degrees in elevation."

In the past, Rotman, lenses were implemented with microstrip or stripline technology, usually between 6 and 18 GHz. Microstrips, however, are unreliable at high frequencies and are therefore not suitable for use in the millimeter-wave region. To reduce the losses to an acceptable value (between 1 and 2 dB), waveguides and air dielectric must be used between the parallel plates in the antenna.

Rausch and Andrew F. Peterson, of the Georgia Tech School of Electrical and Computer Engineering, fabricated the Rotman-lens MMW antenna out of a solid block of aluminum using designs that

BUILD THIS AUDIO EXPANDER

PHILL HAUSMAN

IT SEEMS THAT EVERYBODY IS AN audio enthusiast these days. With compact discs, stereo TV, hi-fi video tapes, and laserdiscs cheaply available to the general public, nobody wants to settle for ordinary sound. If you build our audio processor, you won't have to settle for ordinary sound either! Our audio processor, which can be built in one evening, can be installed easily between any audio output source and any audio input source, and it can "supercharge" any mono or stereo audio signal.

Some of the many uses for the audio processor include enhancing the audio output from a stereo system or any scanner, shortwave, CB, or ham radio receiver. Stereo and mono TV sound can also be greatly enhanced; a pseudo-stereo effect can be added to a mono soundtrack, and if stereo audio is already in use, you can add enhanced stereo effects. The processor also has amazing effects on old mono or pre-Dolby cassette and 8-track tapes—it's great for doing restoration of old recordings.

Features

The audio processor provides a complete array of controls that give the user maximum flexibility in adjusting the processor. The unit has switch-selectable stereo sound from a stereo source, spatial (widened) sound from a stereo source, and pseudo-stereo sound from a mono source; LED's indicate which mode is selected.

The unit exhibits very low noise (-64 dB minimum) and low total harmonic distortion (maximum THD is 0.5% at 1000 Hz). The input and output resistance is 100 kilohms, which is the standard line-level resistance. The left and right channels provide a minimum of 60 dB separation, and there's an adjustable input attenuator for



***Enhance stereo and mono audio,
for under \$30, with our
one-chip audio processor.***

each channel, which provides a minimum of 70.0 dB attenuation. There's independent gain adjustment of the left and right channels in the spatial stereo mode, and independent gain adjustment of the left and right channels in the mono pseudo-stereo mode.

Theory of operation

The audio processor, whose schematic is shown in Fig. 1, is based on the Signetics/Philips TDA3810N stereo, spatial, pseudo-stereo processor IC. A block diagram of the chip is shown in Fig. 2. The device is essentially a system of internal op-amps used as active filters. The chip contains three op-amp stages, switching circuitry, power-supply regulation circuitry, and LED drivers.

The easiest way to understand how the chip works is to follow the audio signal through the chip for each individual mode.

Stereo mode

When the stereo mode is selected, the left input signal

passes through input-attenuator R14, and is coupled via C4 to pin 2 of IC1 (the TDA3810N). The right input signal passes through R15 and C3 to pin 17 of IC1. The various active filters within IC1 serve as a unity-gain bandpass filter with a -3 dB low-frequency cutoff occurring at 20 hertz, and the -3 dB high-frequency cutoff occurring well beyond 20,000 hertz.

Spatial mode

As shown in Fig. 3, when the spatial, or widened, stereo mode is selected, active op-amp circuits selectively remove certain parts of the audio spectrum. The effect is achieved by feeding the audio input to the non-inverting input of an op-amp. The inverting input is fed by the output of the internal buffer/amplifier coupled with a cross-channel signal through R2. Both channels operate in the same way. The gain of each channel can be adjusted independently: R3 adjusts the left channel and R1 adjusts the right channel.

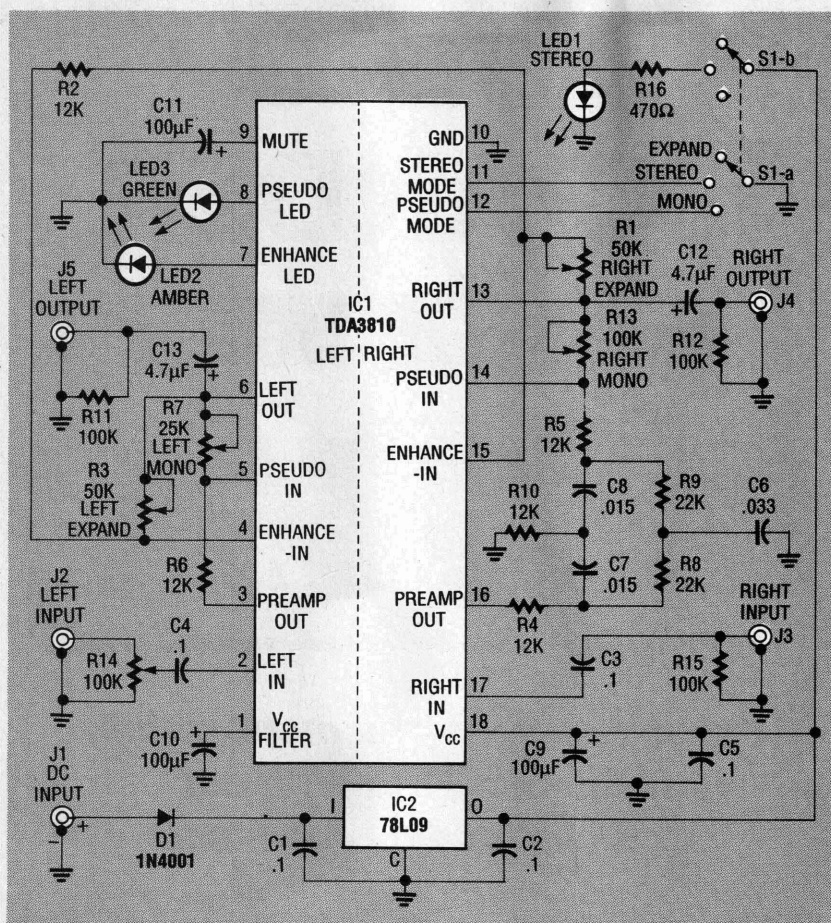


FIG. 1—THE AUDIO PROCESSOR is based on the Signetics/Philips TDA3810N stereo, spatial, pseudo-stereo processor IC.

PARTS LIST

All resistors are 1/4-watt, 5%, unless otherwise noted.

R1, R3—50,000 ohms, panel-mount potentiometer
R2, R4—R6, R10—12,000 ohms
R7—25,000 ohms, panel-mount potentiometer
R8, R9—22,000 ohms
R11, R12—100,000 ohms
R13—R15—100,000 ohms, panel-mount potentiometer
R16—470 ohms

Capacitors

C1—C5—.01 µF, 50 volts, metal film
C6—.0033 µF, 50 volts, metal film
C7, C8—.015 µF, 50 volts, metal film
C9—C11—100 µF, 16 volts, electrolytic
C12, C13—4.7 µF, 35 volts, electrolytic

Semiconductors

IC1—TDA3810N audio processor chip (Philips)
IC2—78L09 9-volt regulator
D1—1N4001 diode
LED1—LED3—light-emitting diodes (use three different colors)

Other components

S1—DP3T panel-mount switch
J1—2.1 mm panel-mount power jack (optional)
J2—J5—panel-mount RCA jack
Miscellaneous: 18-pin IC socket, metal project case, 12-volt DC wall transformer (100-mA), 6 knobs, PC board, shielded cable, wire, solder, etc.

Note: The following items are available from HESC, PO Box 12649, Fort Wayne, IN 46864-2649:

- A kit of parts including a PC board and all parts that mount on it (does not include potentiometers, project case, or wall transformer)—\$19.95 + \$3.05 S&H
 - PC board and TDA3810N—\$14.95 + \$3.05 S&H
 - TDA3810N—\$7.00 postage paid
- Please allow 6 to 8 weeks for delivery.

Pseudo stereo mode

As shown in Fig. 4, when the mono pseudo-stereo mode is selected, the processor creates a "stereo" image using the various active filters differently on each channel. The left channel is fed through a flat-response amplifier (200 Hz to 20 kHz) to achieve a constant gain across the audio spectrum. The gain is adjustable via potentiometer R7, which controls feedback. The optimum setting recommended by Philips, the chip manufacturer, is +2.00 dB at 15 kHz; R7 gives an adjustment range of -69.0 dB to +6.0 dB at 15 kHz.

The right channel uses a twin-T notch filter to give the illusion of signal separation, or the pseudo-stereo effect. Between 100 and 1600 hertz, the notch filter attenuates the audio range by -33.0 dB at its lowest point, which occurs at about 450 hertz. The gain is adjustable by R13, which controls feedback. Philips recommends 0 dB at 15,000 Hz, although R13 gives a range of -68.0 dB to +6.0 dB. Figure 5 shows the frequency response of each channel.

Remaining circuitry

The TDA3810 contains a muting circuit with built-in hysteresis (controlled by C11) that prevents the status LED's from flickering when switching from mode to mode. Also included is the mode selection switch (S1-a and -b) and the corresponding status indicators LED1, LED2, and LED3.

The project requires an external power supply of +10.5–24.0 volts DC, which passes through polarity-protection diode D1 to the input of IC2, a 78L09 +9.0 volt DC, 0.1-amp voltage regulator. Capacitors C1 and C2 provide decoupling and anti-oscillation protection for IC2.

Construction

Most of the parts for the audio processor are available from many electronics distributors. A foil pattern is provided if you want to make your own PC board, or you can use the board available from the source men-

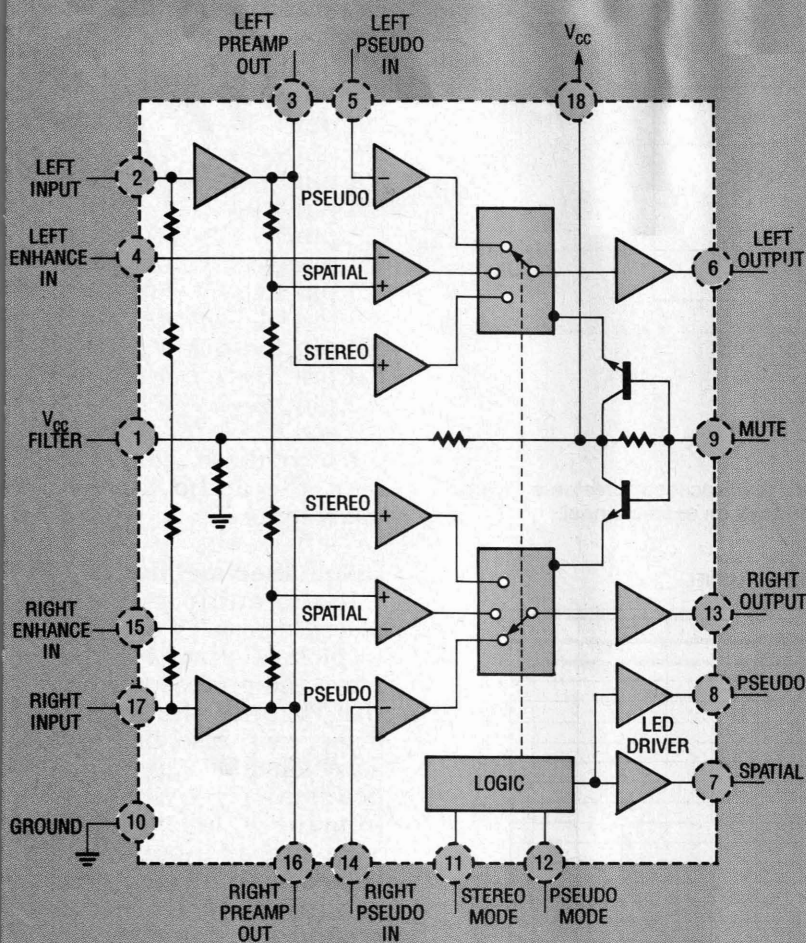


FIG. 2—BLOCK DIAGRAM OF THE TDA3810N. The device contains three op-amp stages, switching circuitry, power-supply regulation circuitry, and LED drivers.

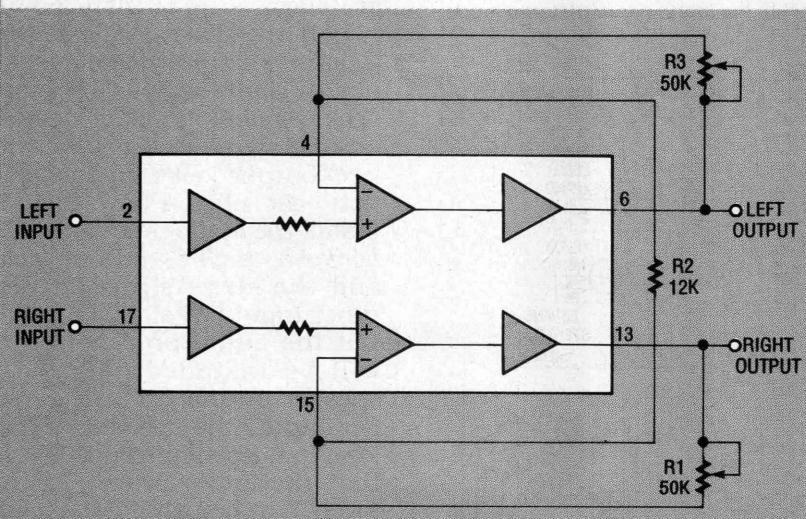


FIG. 3—IN THE SPATIAL STEREO MODE, active op-amp circuits selectively remove certain parts of the audio spectrum.

tioned in the Parts List.

Because this is an audio project, component tolerances and the matching of component values between the left and right channels is important. All re-

sistors should be within 5 percent of the listed value, and it's a good idea to check them. Be sure to use high-quality capacitors, as poor-quality capacitors can adversely effect the overall

frequency response of the circuit's processor.

Following the parts-placement diagram in Fig. 6 as a guide, install all resistors and capacitors, paying attention to the polarity of the electrolytics. Also mount diode D1, observing its polarity. Mount the 78L09 voltage regulator (IC2) with the flat side pointing toward the outside edge of the PC board. It's a good idea to use a socket for IC1, and you can install it now and then insert the IC. Observe the proper pin-1 orientation when putting the chip in its socket.

A metal enclosure will provide the best shielding from interference. However, a plastic box will probably be acceptable as well. Prepare the enclosure by laying out the front and rear panels for drilling. After drilling, deburr and clean the enclosure. The author mounted solder posts at each point on the board that must be hard-wired to another component. That made it easy to mount the board, jacks, and controls in the case, and then do the point-to-point wiring. Otherwise, all wires will have to be soldered to the board before mounting it in the enclosure. In either case, mount the completed PC board to the enclosure using appropriate hardware.

If you haven't done so yet, mount all of the potentiometers, jacks J1-J5, the LED's, and switch S1 to the enclosure. Then solder all cables and wires from the board to the jacks and controls. Following Fig. 6, pay careful attention to the places where shielded cable must be used. Hook-up wire can be used for all other connections. When connecting the feedback gain-control potentiometers (R1, R3, R7, and R13), wire them so their resistance *increases* as they are turned clockwise. Figure 7 shows the inside of the completed prototype.

Testing

When the project is complete, check all components for proper location and polarity, and also check your soldering. Using a DC voltmeter, check for the fol-

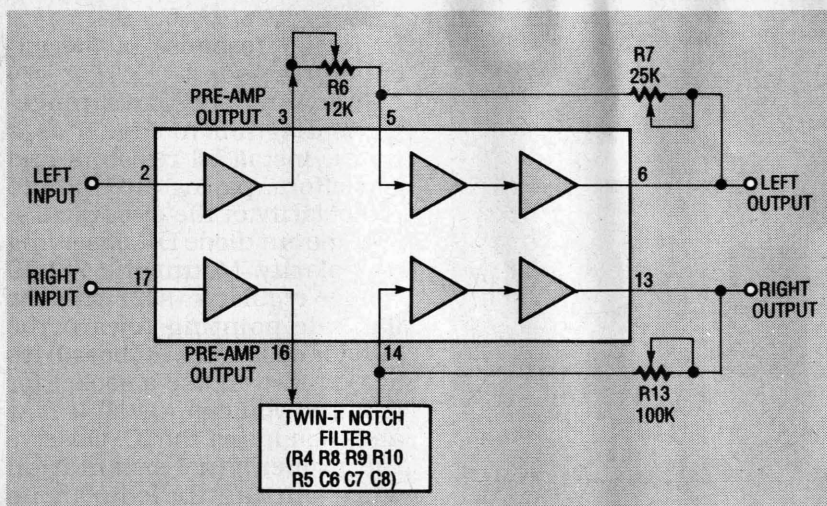


FIG. 4—IN THE MONO PSEUDO-STEREO MODE, the processor creates a "stereo" image using the various active filters in different ways on each channel.

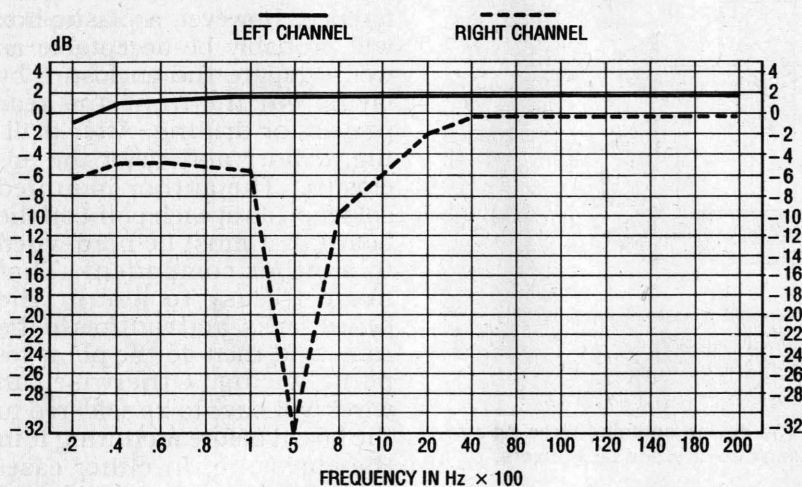


FIG. 5—FREQUENCY RESPONSE of each channel in the mono pseudo-stereo mode.

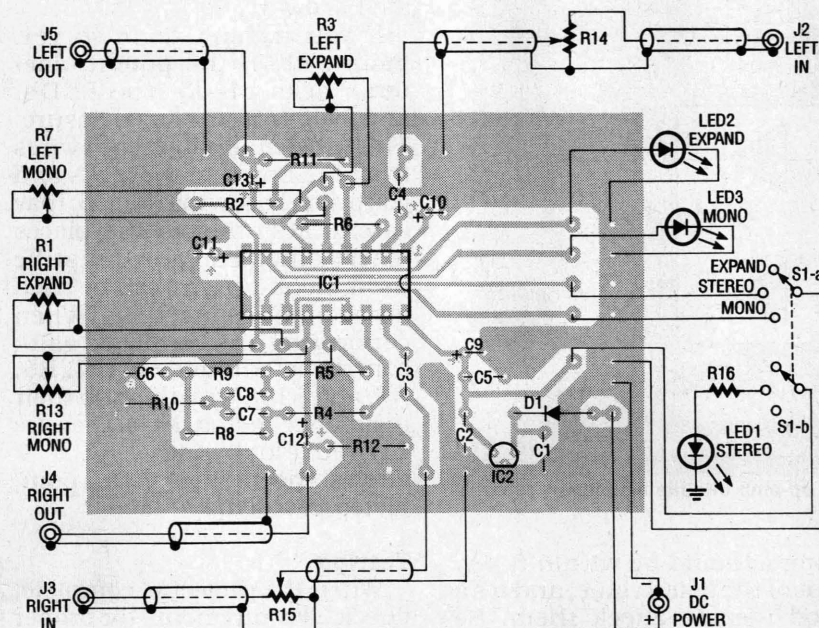


FIG. 6—PARTS-PLACEMENT DIAGRAM. Pay attention to those controls that require shielded cable for connection to the PC board.

lowing voltages after applying power to the circuit:

Power-supply output—10.5—2 VDC

IC2 output—8.55—9.45 VDC

IC1 pin 1—4.2—4.8 VDC

IC1 pin 2—4.2—4.8 VDC

IC1 pin 3—4.2—4.8 VDC

IC1 pin 4—4.2—4.8 VDC

IC1 pin 5—4.2—4.8 VDC

IC1 pin 6—4.2—4.8 VDC

IC1 pin 13—4.2—4.8 VDC

IC1 pin 14—4.2—4.8 VDC

IC1 pin 15—4.2—4.8 VDC

IC1 pin 16—4.2—4.8 VDC

IC1 pin 17—4.2—4.8 VDC

IC1 pin 18—8.55—9.45 VDC

If all of those voltages are correct, the audio processor is ready to use.

Installation and use

If the audio processor is placed near a TV set and tends to pick up stray interference, move the processor around of the TV until the interference disappears. Also, be sure to use a well-filtered DC power pack or power supply to eliminate hum from the AC line.

Be sure all input and output cables are shielded, and well grounded to the PC board ground foil. Improper grounding can cause ground loops and other noise problems that are hard to track down. When the processor is to be installed between the TAPE IN/TAPE OUT jacks on a stereo system, hook it up the same way as if a tape deck were being installed.

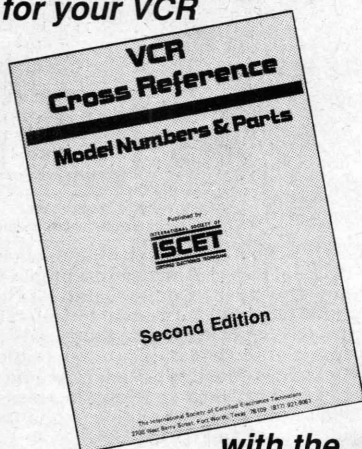
Some stereo TV sets have line level output jacks on them. In that case, all you have to do is install the audio processor unit between the TV's output jack and the stereo's auxiliary or tuner input jacks.

If the audio processor unit will be installed between speaker output and an amplifier input, the speaker output level must be reduced to a level that the audio processor can safely handle, which is a maximum of 2 volts rms. Always set the input-attenuator potentiometers to maximum attenuation before applying an input signal to the processor. Connect all mono audio sources to both of the inputs.

continued on page 82

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AUDIO EXPANDER

continued from page 74

In the stereo mode, the only adjustments are the input-attenuator controls (R14 for the left and R15 for the right). For a line-level input, set them for 0 attenuation. Other higher-level sources should be adjusted as desired by the user (without exceeding the 2-volt rms maximum input limit).

In the spatial stereo mode, the

input attenuator control should be adjusted the same way as in the stereo mode. The spatial controls (R3 for the left and R1 for the right) can then produce very interesting spatial effects. Also, adjusting the input-attenuator controls produces interesting effects due to the cross-channel coupling provided by resistor R2. Experiment with those four controls to determine their optimum settings.

In the mono pseudo-stereo mode, the input-attenuator

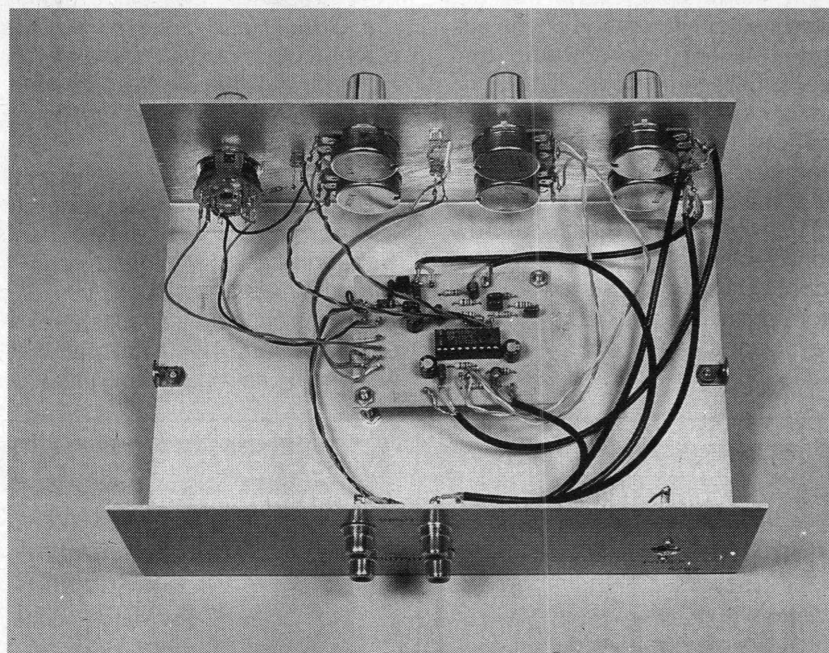
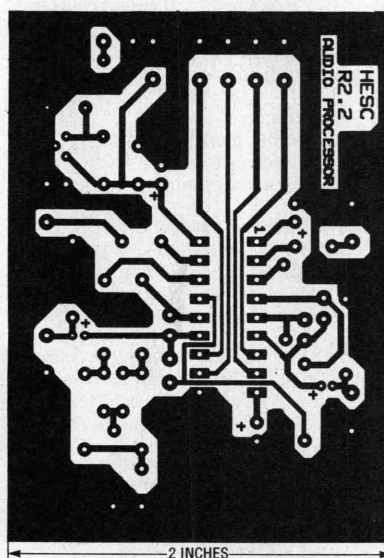


FIG. 7—A METAL ENCLOSURE provides the best shielding from interference. Here's the inside of the completed prototype.



FOIL PATTERN for the audio processor.

controls should be adjusted the same way as in the stereo mode. The pseudo stereo controls (R7 for the left and R13 for the right) should be set as follows: R7 should be set to give a +2.0 dB higher output level versus the right output channel control (R13). For example, set R7 for an output of +2.0 dB, and set R13 for an output of 0 dB. The specific levels are not critical; just keep a 2.0 dB differential between the left and right output channels.

When using a mono input source, the spatial stereo mode will produce some very interesting and useful effects. The input attenuator controls will also have interesting effects.

ew publications for this month. The first is Dick Oliver's new *Nonlinear Nonsense*, a free quarterly newsletter on fractal modeling, chaos science, artificial life, and creative graphics. His 3-D fractal design software is also being offered commercially.

Also free on professional request is the new *Spread Spectrum Handbook* from *Stanford Telecomm*. This is a dozen pages of ap-notes and useful references, along with bunches of data sheets for the company's high-end digital spread-spectrum chips.

The *Scrambling News* has detailed insider information on the current TV video scrambling and descrambling methods. It also offers books full of schematics, modifications, and related technical information. A competing newsletter is the *Satellite Watch News*.

The *SRS Encoder* is a new monthly newsletter from the *Seattle Robotics Society*. Lots of regional information on great surplus buys are included here.

Joe Singer's *Printer's Devil* is a

superb alternative publication on home printing and presswork. "A press in every home; a home in every press." It is published quarterly at a bargain \$10 per year subscription price.

Finally, the *New Age Alternatives Discount Catalog* is another resource for pseudoscience devices and information. Fascinating reading for sure. This one is offered by *Lor'd Industries*.

A reminder here: I have reprints available for most of my columns. Both in *Book-on-demand* published hard copy and on-line at *GENIE PSRT*. The titles now include my *Ask the Guru*, *Hardware Hacker*, the *Blatant Opportunist*, *LaserWriter Secrets*, and the new *Resource Bin*. Check my nearby *Synergetics* ad for more details.

As usual, I have gathered most of the mentioned resources together into our *Names and Numbers* and *Semiconductor Manufacturer's III* sidebars. Also, as usual, you can get technical help, consultant referrals, and lots of off-the-wall networking by calling my helpline. Ω

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FIELD-EFFECT TRANSISTORS (FET'S) are unipolar rather than bipolar devices, and this gives them certain properties that are superior to those of bipolar transistors. Unlike the bipolar transistor, whose current depends on the movement of both electrons and holes, FET operation depends on only one of those charge carriers. Freed from the time delays that occur when those charge carrier recombine, FET's offer faster switching speed and higher cutoff frequencies.

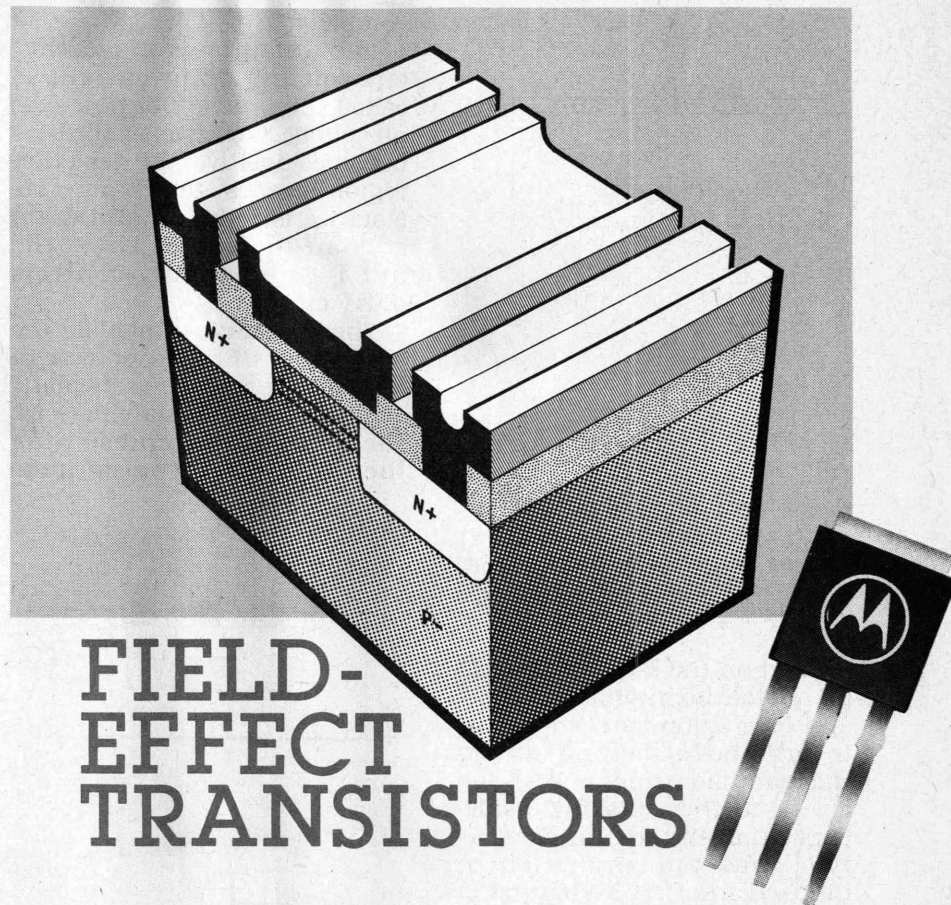
Other advantages of the FET include:

- Voltage rather than current operation.
- Extremely high input impedance in the OFF state.
- Virtually constant current with respect to voltage at specific bias levels.
- Current change that is inversely rather than directly proportional to temperature.

Despite these advantages, the FET has not replaced the bipolar the bipolar transistor in all applications, but it has encouraged new generations of small-signal and general purpose and RF MOSFET's as well as general purpose and RF MOSFET power transistors. Moreover, the latest digital logic families are based on FET technologies. The basic FET is a simple, three-terminal, voltage-controlled device with characteristics that are similar to those of vacuum-tube pentodes. Thus, the FET is considered to be the solid-state equivalent of a pentode.

The two major classes of FET's are the *junction FET* (JFET) and *metal-oxide-semiconductor FET* (MOSFET), formerly called an *insulated-gate FET* (IGFET). FET's are further divided into *N-channel*, *P-channel*, *depletion-mode* and *enhancement-mode* devices. MOSFET's with N-doped channels are called NMOS, and those with P-doped channels are referred to as PMOS.

The three electrodes in all FET's are the *source*, *drain*, and *gate*, analogous to the emitter,



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collector, and base of the bipolar transistor. Both JFET's and MOSFET's are available as discrete transistors. Some MOSFET's have dual gates and are intended for use as radio-frequency mixers.

This article explains how P-channel and N-channel MOSFET's are combined to form the popular complementary-MOS (CMOS) digital logic families. CMOS technology has made possible very large scale integrated (VLSI) memories, microprocessors and dedicated circuits that contain more than one million transistors and still have very low power requirements.

Most small-signal FET's today

are fabricated with *planar* geometry in which all electrodes are accessible from the top of the device. The active regions are defined on the wafer or substrate by successive masking, etching, and deposition or ion-implantation steps. But power MOSFET's now being fabricated with vertical structures are actually capable of handling much higher current in smaller areas of silicon.

Junction FET's

The simplest FET, the JFET, is illustrated by the cross-section view of Fig. 1-a. It is made by selectively implanting or diffusing ions into the wafer or substrate. An N-type region is

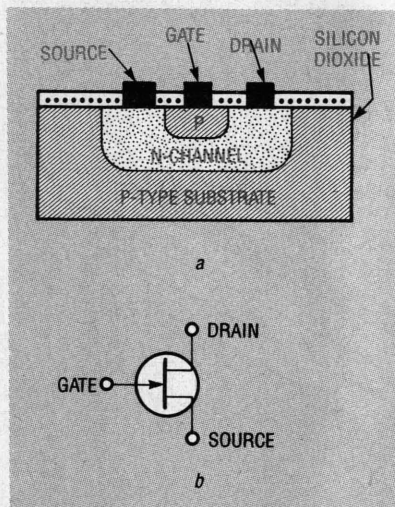


FIG. 1—N-CHANNEL PLANAR JFET showing diffused or implanted channel and gate regions (a), and schematic symbol (b).

defined on the P-type substrate by photolithographic methods, and N-type ions are implanted to form the N-channel. Later in the manufacturing process, following further masking, oxide-deposition and etching steps, P-type ions are implanted or diffused into the N-channel to form the P-type gate.

Aluminum source and drain terminals are formed directly on the N-channel and an aluminum gate terminal is formed on the P-type gate. The symmetrical construction of the JFET permits the drain and source to be interchanged, if necessary.

If a positive voltage is applied at the drain of the N-channel JFET shown in Fig. 1-a, and a negative voltage is applied at the source with the gate terminal open, a drain current flows. When the gate is biased negative with respect to the source, the PN junction is reverse biased, and a *depletion region*, devoid of current carriers, is formed.

Because the N-channel is more lightly doped than the P-type gate material, the depletion region penetrates into the N-channel. This region, depleted of charge carriers, behaves like an insulator. The depletion region narrows the N-channel and increases its resistance. If the gate bias is made even more negative, drain current is cut off

completely.

The gate-bias voltage that cuts off the drain current is called the *pinch-off* or *gate-cut-off voltage*. However, as the bias becomes positive, the depletion region recedes, the channel resistance is reduced, and drain current increases. Thus, the JFET gate actually controls JFET current.

The schematic symbol for the N-channel JFET is shown in Fig. 1-b. As in other schematic symbols for solid-state devices, the arrowhead (representing the direction of conventional

current flow) points from P-doped material to N-doped material. In the N-channel JFET symbol, the arrowhead points from the P-type gate toward the N-type channel.

A section view of a P-channel JFET is shown in Fig. 2-a. The channel of the device is P-type material, and the gate is N-type. If a positive voltage is applied to the source, conventional current flows from the source to the drain. To reverse bias the junction between the N-type gate and the P-type channel, the gate must be made positive with respect to the channel. The biasing voltages of a P-channel JFET are opposite to those of the N-channel JFET.

The schematic symbol for the P-channel JFET is shown in Fig. 2-b. The arrowhead also points from P-type material to N-type material. In this instance, it points from the P-type channel to the N-type gate region. The characteristics of the P-channel JFET are similar to those of the N-channel device, except that the voltage and current polarities are reversed.

Both N-type and P-type JFET's operate in the *depletion mode*; that is, they conduct with zero bias on their gates. Figure 3 shows a typical family of *drain characteristics* for an N-channel JFET. As the gate-to-

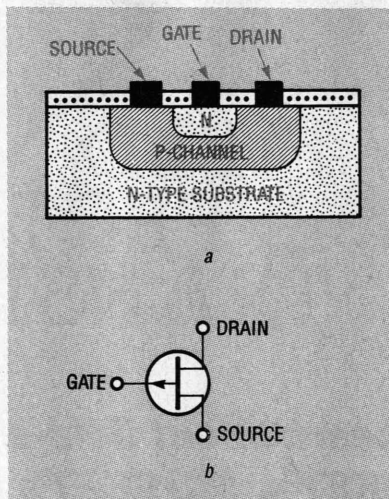


FIG. 2—P-CHANNEL PLANAR JFET showing diffused or implanted channel and gate regions (a), and schematic symbol (b).

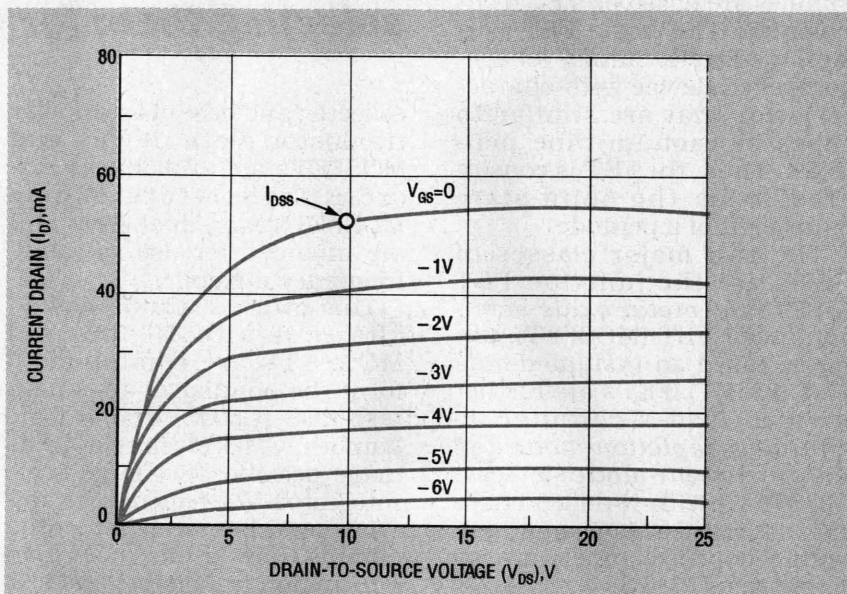


FIG. 3—DRAIN CHARACTERISTICS FOR AN N-CHANNEL JFET. The family of gate curves has its origin at zero drain-to-source volts, and the gate bias values show voltage control.

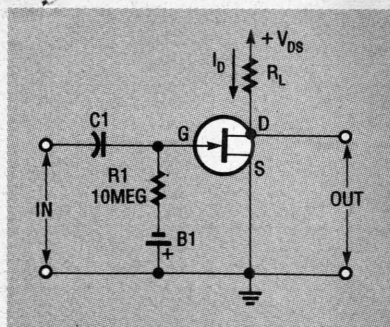


FIG. 4—N-CHANNEL JFET COMMON-source amplifier is analogous to a bipolar common-emitter amplifier.

source voltage is made increasingly negative, the depletion region is increased, and drain current decreases. As a result, pinchoff voltage occurs at a lower value of V_{DS} . Curves for different values of gate-to-source bias, V_{GS} , are plotted in the figure because the FET is a voltage-operated device.

JFET circuits

When an N-channel JFET is connected to a V_{DS} supply as shown in Fig. 4, a drain current, I_D , flows in the device. The magnitude of I_D can be controlled by a gate-to-source bias voltage, V_{GS} . Similarly, when a P-channel JFET is connected to a negative drain voltage, a drain current, I_D , flows in the device. The value of I_D is maximum when V_{GS} equals zero, and it is reduced (to bring the JFET into a linear operating region) by applying a reverse bias to the gate terminal of the device (negative bias in a N-channel device, positive bias in a P-type).

In Fig. 3, the value of V_{GS} to reduce I_D to zero, the gate-to-source pinchoff voltage V_P is about -7 volts. The value of I_D when V_{GS} equals zero (called I_{DSS} or drain saturation current for zero bias) is about 52 milliamperes for the device shown in the figure.

The gate-to-source junction of the JFET has the characteristics of a silicon diode. When reverse biased (to bring it into its linear operating region), gate leakage currents (I_{GSS}) are measured in thousandths of a microampere at room temperature. Actual gate signal currents are only a fraction of a

that, and the input impedance to the gate is typically 1000 megohms at low frequencies. The gate junction is effectively shunted by a capacitance of a few picofarads, so input impedance falls as input frequency is increased.

If the gate-to-source junction of the JFET is forward biased, it conducts like a normal silicon diode, and if it is severely reverse biased it avalanches like a Zener diode. Neither of those conditions will harm a JFET if its gate currents are limited to those specified.

Referring to the N-channel JFET drain characteristics in Fig. 3, it can be seen that, for each value of V_{GS} , drain current I_D rises linearly from zero as the

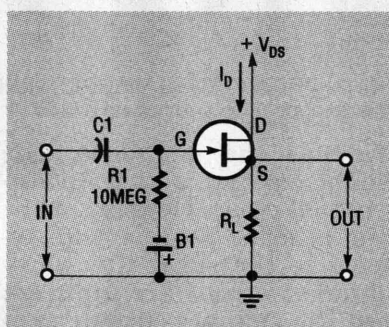


FIG. 5—N-CHANNEL JFET COMMON-drain (source-follower) amplifier is analogous to a bipolar emitter-follower amplifier.

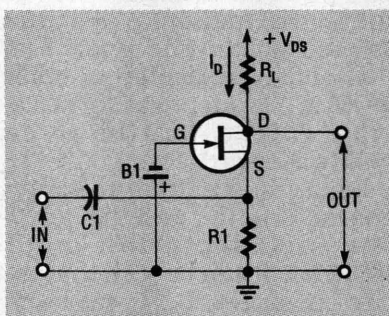


FIG. 6—N-CHANNEL JFET COMMON-gate amplifier is analogous to a bipolar common-base amplifier.

drain-to-source voltage (V_{DS}) is increased from zero to a value at which a knee occurs on each curve. Moreover, I_D remains virtually constant as V_{DS} is increased beyond where the knee occurs.

Thus, when V_{DS} for any of the family of V_{GS} curves is below its knee value, the drain-to-source

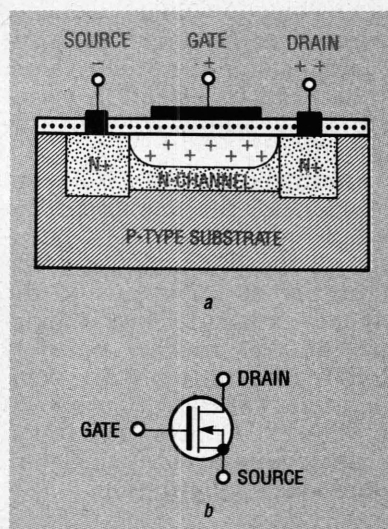


FIG. 7—N-CHANNEL, DEPLETION-mode MOSFET with negative gate bias (a), and schematic symbol (b).

pins of the JFET act like a voltage-variable resistor with its value determined by V_{GS} . The drain-to-source resistance, R_{DS} , can be varied from several hundred ohms at $V_{GS} =$ zero to thousands of megohms at pinchoff. That characteristic permits the JFET to be used in a circuit as a voltage-controlled switch.

From the drain characteristic curve of Fig. 3, it can be seen that when V_{DS} is above the knee value, the I_D value is dictated primarily by the V_{GS} value, and is virtually independent of the V_{DS} value. This characteristic permits the JFET to function as a voltage-controlled current generator.

The gain of a JFET is specified as a *transconductance*, g_m , the rate of change of drain current with respect to gate voltage. A g_m of 5 milliamperes per volt indicates that a variation of one volt on the gate produces a change of 5 milliamperes I_D . The units of this measurement are in inverse ohms or *mhos*. You will find that JFET data sheets usually specify g_m in millimhos or micro-mhos.

The N-channel JFET in Fig. 4 is organized as a common-source amplifier, analogous to a bipolar NPN common-emitter amplifier. In typical applications, the JFET is biased into its linear region and organized

as a voltage-to-voltage converter or amplifier. As shown in Fig. 4, a load resistor of suitable value, R_L , should be placed in series with the JFET's drain-to-source current.

Another common JFET configuration is the *common drain* or *source-follower* configuration shown in Fig. 5. That configuration is analogous to the bipolar emitter-follower configuration. Yet another possible JFET configuration is the *common-gate* configuration shown in Figure 6. That configuration is analogous to a bipolar common-base configuration.

MOSFET's explained

The metal-oxide FET or MOSFET was developed as an improvement on the JFET, and it has become the most important form of FET. Figure 7-a illustrates an N-channel depletion-mode MOSFET with a negative gate bias. The gate of this MOSFET is fully insulated from the adjacent channel. This is the most important distinction between an N-type depletion-mode MOSFET and an N-type JFET, which is manufactured with a doped gate region directly under and in contact with the gate.

The surface of the silicon P-type wafer is first coated with a layer of silicon dioxide (SiO_2), and the source and drain windows are masked and etched to expose the P-type substrate. N-dopants are heavily diffused or implanted into those two regions. Another window is masked and etched over the channel, and it is given a lighter concentration of N dopant. In subsequent steps, the channel is recoated with an insulating oxide, and the metal source, drain, and gate terminals are deposited.

When the drain is positive with respect to the source, a drain current will flow, even with zero gate voltage. However, if the gate is made negative with respect to the substrate, positive charge carriers (holes) induced in the N-channel will combine with the electrons and cause channel resistance to increase. With increasing nega-

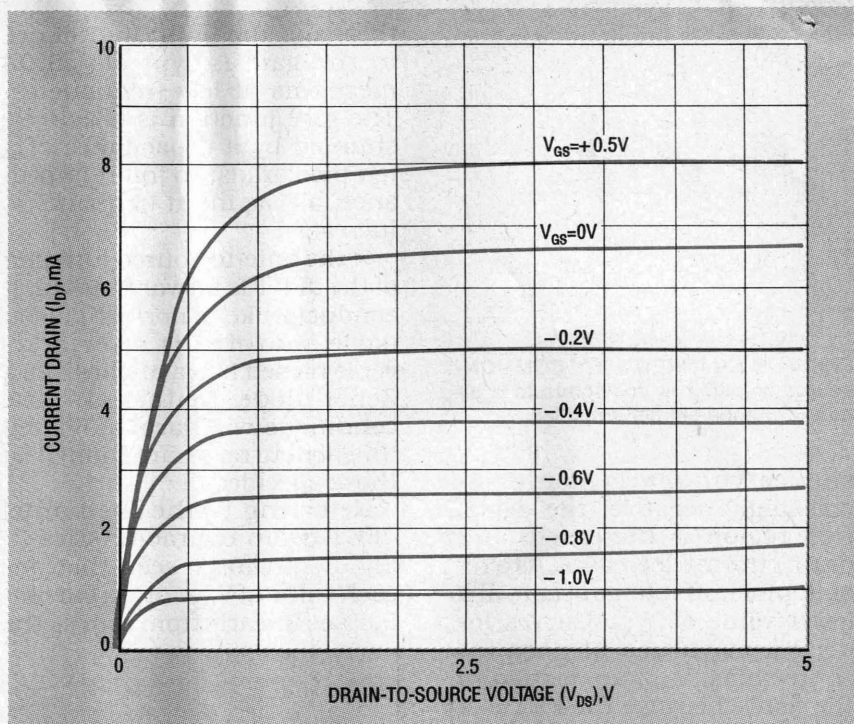


FIG. 8—DRAIN CURVES FOR N-CHANNEL DEPLETION-MODE MOSFET showing the effects of positive and negative bias

tive bias, the pinchoff voltage will be reached, and drain current will cease. However, if the gate is made positive with respect to the substrate, additional electrons are induced, and the channel current then increases.

The schematic symbol for the N-type depletion-mode MOSFET is shown in Fig. 7-b. The path or channel between the source and drain is shown as a solid bar. The symbol for the P-channel depletion-mode MOSFET is identical to the N-type, except that the arrow points outwards.

Figure 8 is a drain-to-source characteristic curve for an N-channel depletion-mode MOSFET. It can be seen that the current drain, I_D , is inversely proportional to the magnitude of the negative gate voltages, V_{GS} . Compare Fig. 8 with Fig. 3 for the N-channel JFET to see their similarities.

Planar *enhancement-mode* MOSFET's are made by the same methods as planar depletion-mode MOSFET's. However, the N-channel enhancement-mode MOSFET shown in Fig. 9 does not have the N-doped drain-to-source channel through the P-type substrate of

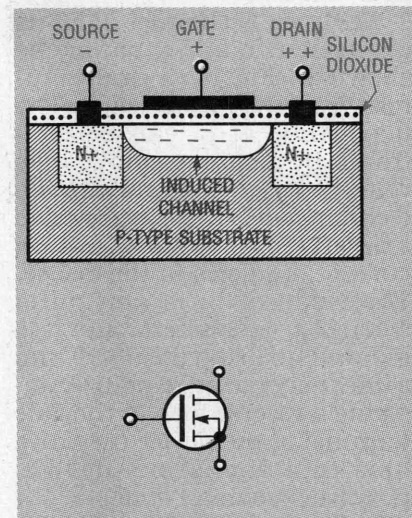


FIG. 9—N-CHANNEL ENHANCEMENT-mode MOSFET with positive grid bias (a), and schematic symbol (b).

the N-channel depletion-mode MOSFET. Therefore, there is no conduction between drain and source at zero gate bias.

To turn an enhancement-mode MOSFET on, positive gate bias is needed. As the gate voltage is increased, more electrons are induced into the channel. They cannot flow across the oxide layer to the gate, so they accumulate at the substrate surface below the gate oxide. When a sufficient number of

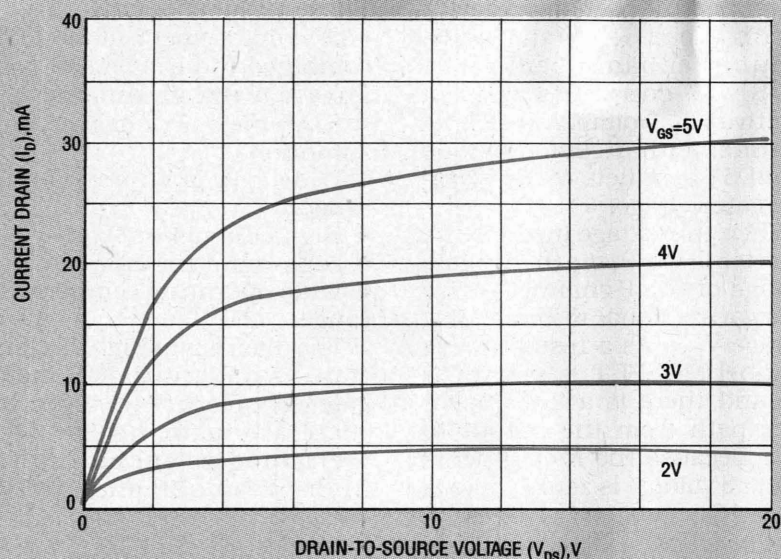


FIG. 10—DRAIN CURVES FOR AN N-CHANNEL, enhancement-mode MOSFET showing the effects of increasingly positive gate bias.

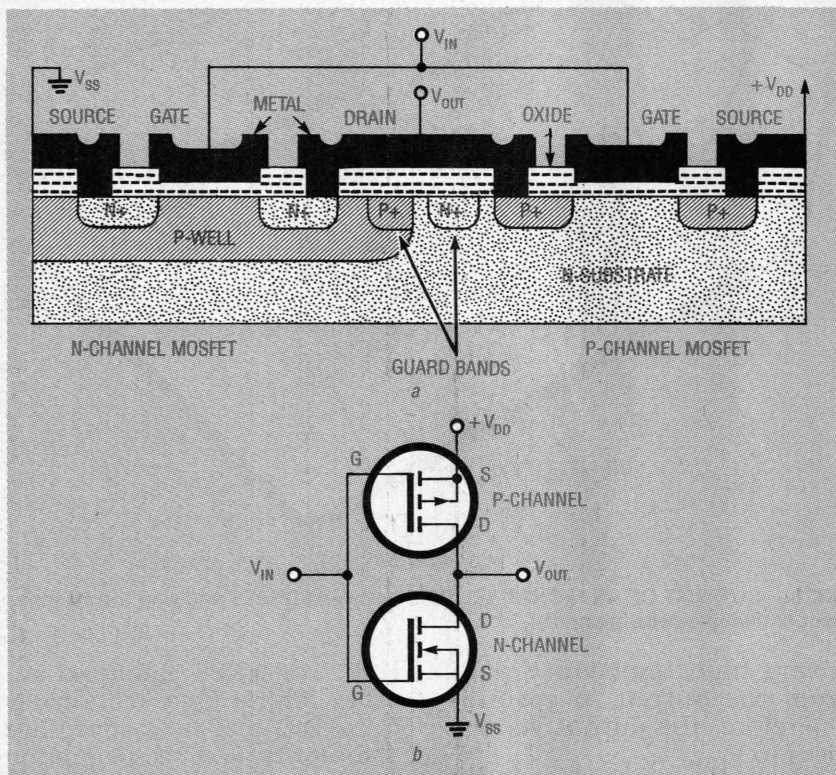


FIG. 11—CMOS INVERTER (NOT) GATE FORMED FROM N-CHANNEL AND P-channel enhancement-mode MOSFET's with power, ground, input, and output connections as shown in (a), and schematic (b).

electrons has accumulated, the P-type substrate material is converted into an N-channel, and drain-to-source conduction occurs. The magnitude of the drain current depends on the channel resistance, but it is controlled by the gate voltage.

The schematic symbol for an

N-type enhancement-mode MOSFET is shown in Fig. 9-b. In this symbol, the gate does not make direct contact with the channel. The arrowhead points toward the (induced) N-type channel, shown as a line broken into three sections to indicate an in-

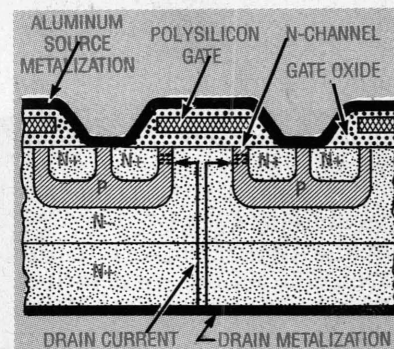


FIG. 12—SECTION VIEW OF AN N-channel DMOS power MOSFET showing its vertical structure and vertical current flow.

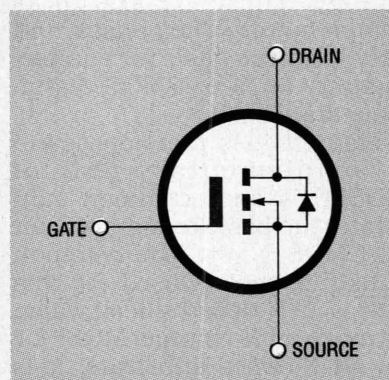


FIG. 13—SCHEMATIC SYMBOL FOR A DMOS power MOSFET including a drain-source diode.

termittent channel.

Current flow in the channels of both kinds of enhancement-mode MOSFET's is proportional to the voltage on their gates, V_{GS} . This can be seen for an N-type enhancement-mode MOSFET by examining the family of gate voltage (V_{GS}) curves in Fig. 10. Current drain, I_D , is directly proportional to the positive value of gate voltage.

A P-channel enhancement-mode MOSFET is made the same way as the N-channel enhancement device except that P-type drain and source regions are diffused into an N-type substrate. The symbol for a P-type enhancement-mode MOSFET is the same as the one shown in Fig. 9-b except that the direction of the arrow is reversed. In the case of a P-type enhancement-mode MOSFET, the drain current is directly proportional to the negative values of its grid voltage.

The high gate impedance of all MOSFET's makes them susceptible to damage from even

low-energy electrostatic discharge (ESD). For this reason many discrete MOSFET's and IC's based on MOSFET's are protected with on-chip Zener diode circuits.

CMOS logic devices

An enhancement-mode MOSFET can act as a switch when it is turned on or off by a voltage applied to the gate electrode: N-channel MOSFET's are switched with positive gate voltage, and P-channel MOSFET's are switched with negative gate voltage. These are known as *complementary responses*, and they form the basis for complementary MOS or CMOS digital logic families.

Figure 11-a is a section view of a complementary pair of MOSFET's on a common substrate, the basic topography for all CMOS gates. The common substrate that is used for this pair is an N-doped silicon wafer. To make an N-channel MOSFET on an N-doped substrate, it is necessary to diffuse or implant a P-doped *well* in the substrate. The smaller N-type wells can then be formed in this P-doped region.

Because the substrate is N-doped, fewer steps are required to form the P-channel FET. The P- and N-doped guard bands isolate and insulate the individual transistors in this integrated circuit to prevent mutual interference. (Although not illustrated here, these guard bands are actually N- or P-doped rings formed around the complete FET below the oxide layer in this CMOS technology.

The two transistors in the section view, Fig. 11-a, can be connected to form a CMOS logic inverter, the simplest of digital logic circuits. This is accomplished by connecting the gates together to form an input (V_{IN}) terminal, and taking the output (V_{OUT}) from the common drain. The source on the left side of the diagram, V_{SS} , is grounded, while the source on the right side is connected to the positive supply, V_{DD} .

Those connections are shown schematically in Fig. 11-b. How does the inverter work? Consid-

er the P-channel device to be the driver and the N-channel device to be the load. Recall that an N-channel enhancement-mode MOSFET conducts with a positive gate voltage, while a P-channel, enhancement-mode MOSFET conducts with a negative gate voltage.

When the voltage input to the inverter is low (logic 0), the gate voltage of the P-channel device is negative, equal to the supply voltage V_{DD} . As a result, the P-channel MOSFET is switched ON, and there is a low impedance path from the output to V_{DD} . Because the N-channel is off (gate voltage is zero), there is

CMOS unit is extremely low. These properties of N- and P-type enhancement-mode FET's combined to form CMOS gates provide many advantages:

- Extremely low power consumption.
- Wide power supply voltage range.
- High DC noise margin
- High input impedance
- Wide operating temperature range.

The diagram in Fig. 11-a illustrates standard CMOS metal-gate technology (74C/4000), but there are many other CMOS technologies including the high-speed silicon-gate HC,

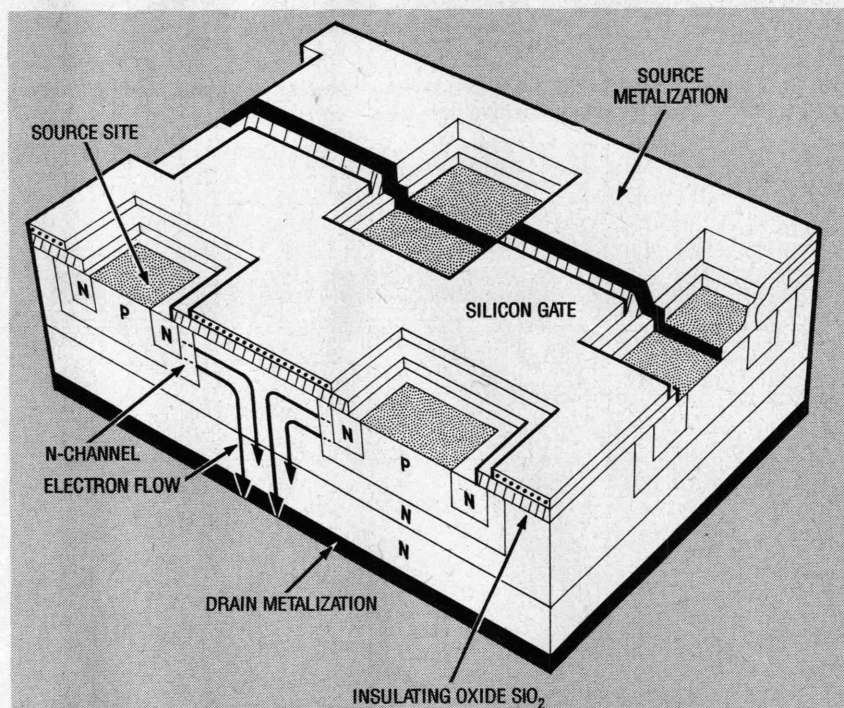


FIG. 14—CUTAWAY OF AN N-CHANNEL DMOS power MOSFET showing the arrangement of multiple cells in parallel.

a very high impedance path from the output to ground. Therefore, the output voltage rises to V_{DD} .

When the input voltage is high (logic 1), the situation is reversed. The P-channel FET is cut off, and the N-channel FET is ON, so the output voltage falls to zero. Therefore, the circuit is a logic inverter: a low input results in a high output, and vice versa.

In either logic state one FET is ON while the other is OFF. Because one FET is always turned off, the quiescent current of the

HCT, and FACT families. Another digital logic technology called BiCMOS takes advantage of the lower power consumption and higher integration density of CMOS, and the higher speed and superior drive capability of bipolar transistors.

Power MOSFET's

Power MOSFET's exhibit the properties of small-signal MOSFET's such as high-input impedance and voltage control, and they have drains, sources and gates, but they are designed

continued on page 89

FET'S

continued from page 56

to handle higher currents. As majority-carrier devices that store no charges, they can switch faster than bipolar power transistors.

Figure 12 is a section view of an N-Channel, enhancement-mode power MOSFET. Unlike its small-signal counterpart, the latest power MOSFETs are fabricated with vertical rather than planar structures. They are made with the double-diffused (DMOS) process, and they have conductive silicon (polysilicon) gates. The gate of this device is isolated from the source by a layer of insulating silicon oxide.

When a voltage is applied between the gate and source terminals, an electric field is set up within the MOSFET. This field alters the resistance between the drain and source terminals, and it permits conventional current to flow in the drain in response to the applied drain circuit voltage. There are also P-channel, enhancement-mode power MOSFETs in which conventional current flows in the opposite direction of the N-channel device. Figure 13 is the schematic symbol for a DMOS enhancement-mode, N-channel power MOSFET.

Figure 14 is an cutaway view of a typical DMOS power MOSFET. It is made up of many cells or transistor elements connected in parallel. Each source cell consists of a closed rectangular or hexagonal channel which separates a source region from the substrate drain body. The cells are formed in an integrated circuit process, and there might be more than a half million cells per square inch of substrate. All of the source cells are connected in parallel by a continuous deposition of aluminum metalization, which forms the grid-like common source terminal.

The DMOS power MOSFET contains an inherent PN junction diode, and its equivalent circuit can be considered as a diode in parallel with the

source-to-drain channel, as shown in the schematic symbol of Fig. 13.

International Rectifier (IR) makes DMOS power MOSFETs that have hexagonally shaped cells, so it calls its products HEXFET's. Motorola Semiconductor also offers DMOS power MOSFETs, but its devices have rectangular rather than hexagonal cells. Motorola named its power MOSFETs TMOS to call attention to the T-shaped current flow that occurs in the cells between the common drain and the channels to the multiple sources.

Power MOSFETs are widely specified for high-frequency switching power supplies (generally those that switch at frequencies above 100 kHz), AC and DC motor speed controls, high-frequency generators for induction heating, ultrasonic generators, audio amplifiers, and amplitude modulation transmitters.

The advantages to using power MOSFETs over power bipolar transistors include:

- Faster switching speeds and lower switching losses.
- Absence of the bipolar's second breakdown
- Wider safe operating area
- Higher input impedance
- High, if not higher, gain
- Faster rise and fall times
- Simple drive circuitry

The principal disadvantages of power MOSFETs are their higher cost and a higher static drain-to-source on-state resistance, which can cause unacceptable power losses in certain switching applications. However, the manufacturers have made progress in reducing those resistance values. DMOS geometry has largely replaced the V-groove or VMOS process that was widely used to fabricate power MOSFETs back in the 1970's.

Radio-frequency power MOSFETs are now available that will operate over the 2 to 200 MHz frequency range. The high power and high gain of these devices makes them suitable as power amplifiers in solid-state transmitters for FM and TV broadcasting. Ω

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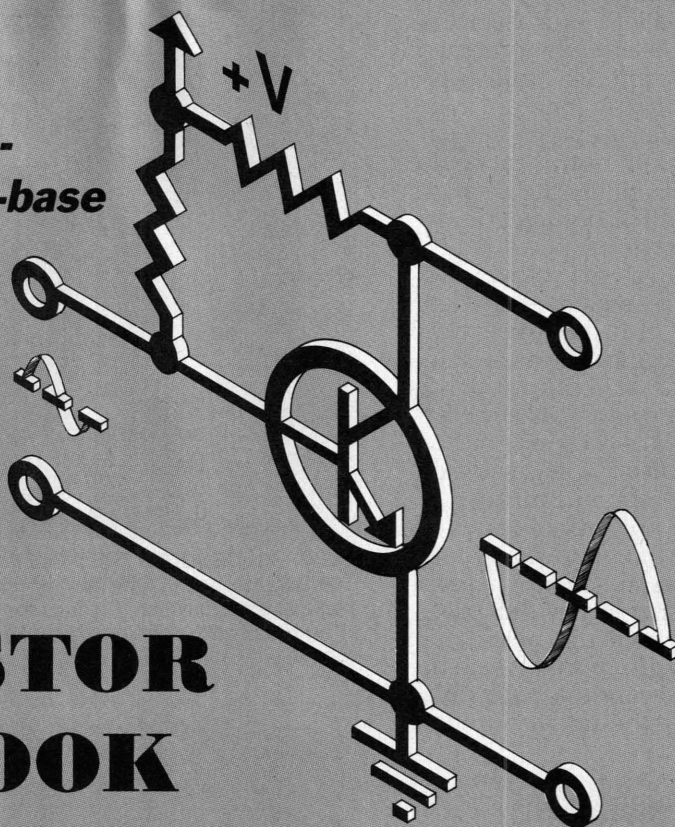
CIRCLE 177 ON FREE INFORMATION CARD

March 1993 Electronics Now

89

Learn about common-emitter and common-base bipolar transistor amplifiers to help you design your own circuits.

TRANSISTOR COOKBOOK



RAY MARSTON

THE CHARACTERISTICS OF COMMON-emitter and common-base bipolar junction transistor (BJT) amplifiers differ from those of the common-collector amplifier—and each other. Those amplifiers are the subjects of this article. Last month the common-collector amplifier was discussed in detail in this series, and many different practical circuits were presented.

Figure 1 shows the three basic bipolar transistor amplifier circuits. It was reprinted from last month's article as was Table 1, which compares the characteristics of the three basic transistor amplifiers. Examination of Table 1 will reveal that the common-collector amplifier (Fig. 1-a) provides near-unity voltage gain, while presenting high input and very low output impedance.

Recall from last month's article that the common-collector amplifier is applied both as a unity-gain *voltage follower* and an impedance converter. By contrast, both the common-emitter (Fig. 1-b) and common-

base amplifier (Fig. 1-c) circuits provide high-voltage gain, so they function primarily as voltage amplifiers. The common-base circuit, for example, offers near unity current gain, so it usually functions as a wide-band or high-frequency voltage amplifier.

The common-emitter circuit, on the other hand, offers both high current gain and high voltage gain, making it an attractive high-gain power amplifier. This circuit, also known as a grounded-emitter circuit, is typically functions as a digital or analog amplifier.

Digital circuitry

Figure 2 is a schematic for a simple NPN common-emitter amplifier that can function as a digital amplifier, inverter, or switch. The input signal can be either zero volts or a high positive value. (It should be higher than 0.6 volt but less than the power supply voltage.) When the input is at zero volts,

the transistor is fully cut off; in that state the output equals the positive power supply voltage.

When the input is switched to a positive value greater than 0.6 volt (the value needed to forward bias the base-emitter junction of a 2N3904 transistor), the transistor turns on. Collector current flows in load resistor R_L and "pulls" the output voltage toward zero.

If the input voltage is high enough, the transistor is driven fully on, or into *saturation*. Then the output voltage falls to a saturation value of only 0.2 to 0.3 volt. As a result, the output signal is an inverted version of the input waveform.

In Fig. 2, resistor R_B acts principally as a protective device to limit the base-drive current to a non-destructive value. The input impedance of the circuit is slightly greater than the R_B value. The value of resistor R_B is inversely related to the waveform rise and fall times: The higher its value, the slower are those times.

This circuit drawback can be

overcome by shunting R_B with a "speed-up" capacitor with a value of about $0.001 \mu\text{F}$, as shown dotted in Fig. 2. In practical applications, R_B should have as low a value as practical, consistent with protecting the transistor and input-impedance requirements. However, it should *never* be greater than $R_L \times h_{FE}$.

Figure 3 is the schematic for a PNP version of the digital inverter or switch circuit. Here the transistor is switched fully ON when the input is zero volts. In that condition, the output is about 0.2 volt less than the positive power supply value. The 2N3906 transistor turns off only when the input rises to a value that is within 0.6 volt of the supply value. The output then falls to zero volts.

The ability of the circuits in Figs. 2 and 3 to respond to lower input signals (sensitivity) can be increased by replacing Q1 with two transistors in a Darlington pair. The circuits in both Figs. 4 and 5 are high-gain, non-inverting digital am-

TABLE 1
CHARACTERISTICS OF THE THREE BASIC TRANSISTOR AMPLIFIERS

Parameters		Common Collector	Common Emitter	Common Base
Input impedance	Z_{IN}	High ($\approx h_{FE} \times R_L$)	Medium ($\approx 1.0K$)	Low ($\approx 40\Omega$)
Output impedance	Z_{OUT}	Very low	$\approx R_L$	$\approx R_L$
Voltage gain	A_V	≈ 1	High	High
Current gain	A_I	$\approx h_{FE}$	$\approx h_{FE}$	≈ 1
Cutoff Frequency	—	Medium	Low	High
Voltage phase shift.	—	0°	180°	0°

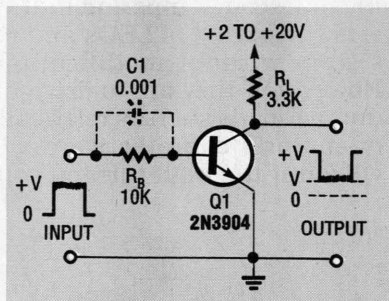


FIG. 2—DIGITAL INVERTER/SWITCH based on an NPN transistor.

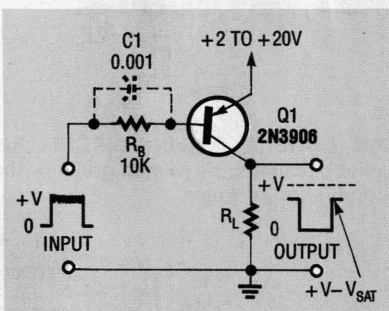


FIG. 3—DIGITAL INVERTER/SWITCH based on a PNP transistor.

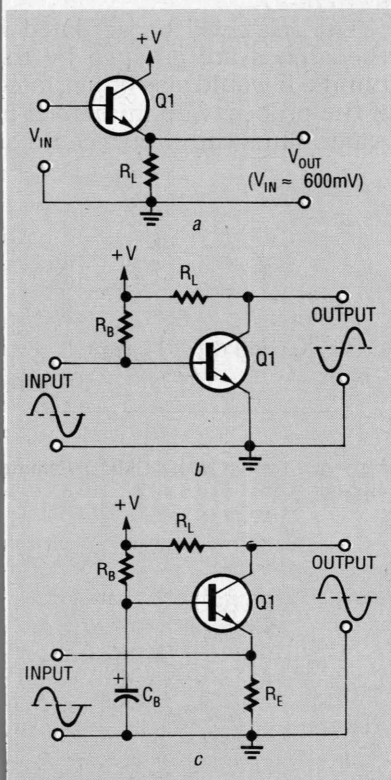


FIG. 1—THREE BASIC TRANSISTOR amplifiers: common-collector (a), common-emitter (b), and common-base (c).

plifiers or digital switches.

The circuit in Fig. 4 which includes the Darlington pair of NPN transistors operates as follows: When the input signal is zero volts, Q1 is cut off, effectively removing it from the circuit. Under this condition, Q2 is driven into saturation through resistor R2, and the output signal has a threshold value of 0.2 to 0.4 volt.

By contrast, if the input signal is significantly greater than 0.6 volt, Q1 is driven into saturation and pulls the base of Q2 down to only about 0.2 volt above zero. Under this condition, Q2 is cut off and the output is at the full supply voltage.

The circuit in Fig. 5, which includes one NPN and one PNP transistor, operates in a dif-

ferent way than that described for Fig. 4. When the input is at zero volts, Q1 is cut off, which, in turn, cuts off Q2 through resistors R2 and R3. Therefore, the output value is zero volts.

However, when the input to the base of Q1 goes high (above 0.6 volt), Q1 is driven on and obtains most of its collector current from the base of Q2 through R3. This action drives Q2 into saturation. When this happens, the output reaches a value about 0.2 volt less than the positive supply voltage.

Figure 6 is a conceptual sche-

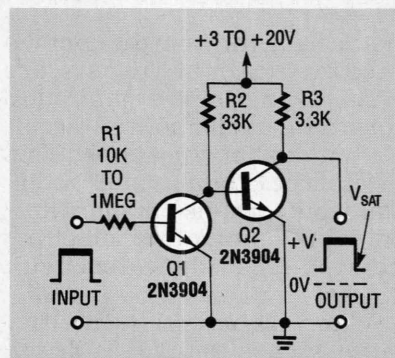


FIG. 4—HIGH-GAIN, NON-INVERTING digital amplifier based on a single NPN transistor.

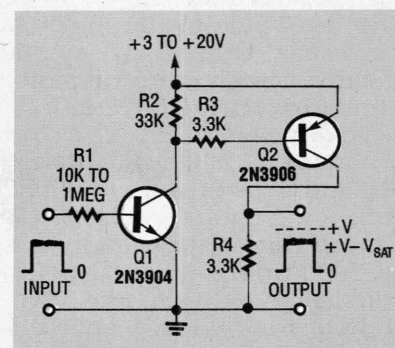


FIG. 5—ALTERNATIVE NON-INVERTING digital amplifier or switch with NPN and PNP transistors.

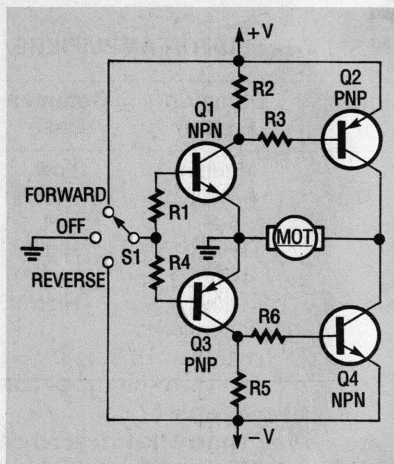


FIG. 6—DC-MOTOR DIRECTION CONTROL circuit shown without component values.

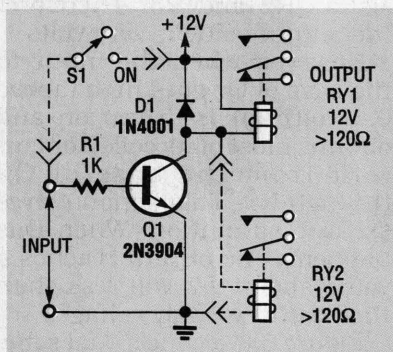


FIG. 7—SIMPLE RELAY-DRIVING circuit.

matic showing how the complementary pair of transistors from Fig. 5 can be organized to form a DC-motor control circuit with a dual power supply. This circuit could be operated by digital logic, and has applications in robotics or other machine control. This is how the circuit works:

When switch S1 is in the FORWARD position, Q1 is driven on through R1, and Q2 is driven on through R3 by Q1. However, Q3 is cut off through R4, and Q4 is cut off by R5 and R6. Thus the "live" side of the motor is connected through Q2 to the positive supply, and the motor runs in a forward direction.

When S1 is in the OFF position, Q1 is cut off through R1, and Q2 is cut off through R2 and R3. Simultaneously Q3 is cut off through R4, and Q4 is cut off through R5 and R6. Under this condition, the "live" side of the motor is open-circuited so the motor does not run.

Finally, when S1 is in the

REVERSE position, Q3 is biased on through R4, and Q4 is driven on through R6 and Q3. However, Q1 is cut off through R1, and Q2 is cut off through R2 and R3. Therefore, the "live" side of the motor is connected through Q4 to the negative power supply, and the motor runs in the reverse direction.

Relay drivers

The basic digital circuits shown in Figs. 3 to 5 can drive a variety of resistive loads such as incandescent lamps and indicators consisting of LEDs and resistors without modification. However, if they are to drive inductive loads such as relay coils or motors, a protective diode will limit the high turn-off volt-

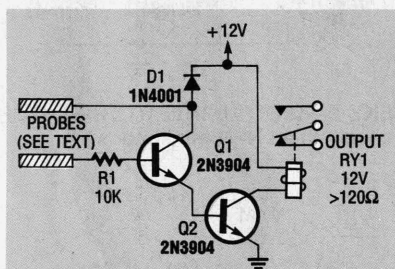


FIG. 8—RELAY-DRIVING CIRCUIT that can be actuated by human grasp or the conduction of water.

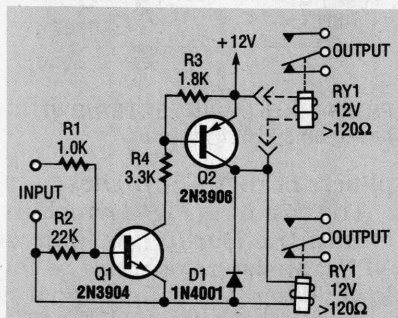


FIG. 9—ULTRA-SENSITIVE RELAY driver needs a 0.7-volt, 40-microampere input.

age to a value that can be handled safely by the transistor.

Common-emitter amplifiers are more sensitive relay drivers than the common-collector amplifiers described last month. Figures 7, 8, and 9 are practical common-emitter relay-driving circuits.

Figure 7, for example, is the schematic for a simple but versatile single-transistor relay driver. It increases the relay's operating current sensitivity by

a factor of about 200 (the nominal h_{FE} value of transistor Q1). Resistor R1, which provides base-drive protection, can have a larger value than the 1 kilohm shown in Fig. 7.

The relay can be turned on either by applying a DC input voltage greater than 0.7 volt or by operating switch S1 (shown as a plug-in component connected by dotted lines).

The circuit shown in Fig. 7 is non-latching, but it can be made self-latching by including another relay (RY2) between the collector and emitter of Q1, as shown in the diagram. The circuit's current sensitivity is limited by the current gain of Q1, but it can be increased to about 20,000 by replacing Q1 with a Darlington pair as shown in Fig. 8.

In a practical application, the relay in the Fig. 8 circuit can be actuated by shorting a pair of stainless-steel probes, each with a resistance value less than a few megohms. Tap water and human skin have resistance values under a few megohms, so this circuit can function as a water, or touch-operated relay switch.

The relay will be actuated if the probes are grasped by the hands. It would also be actuated if the probes were immersed in water. Thus the circuit could be

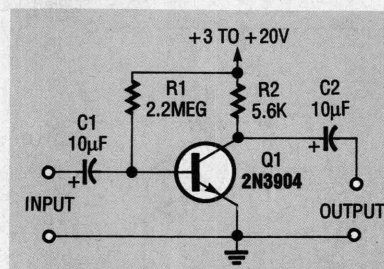


FIG. 10—SIMPLE NPN COMMON-emitter amplifier.

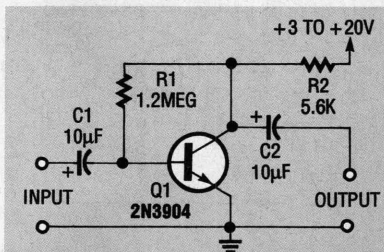


FIG. 11—COMMON-EMITTER amplifier with feedback biasing.

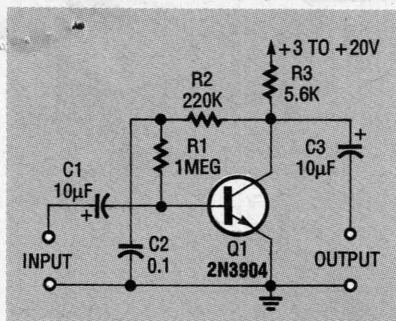


FIG. 12—COMMON-EMITTER amplifier with AC-decoupled feedback biasing.

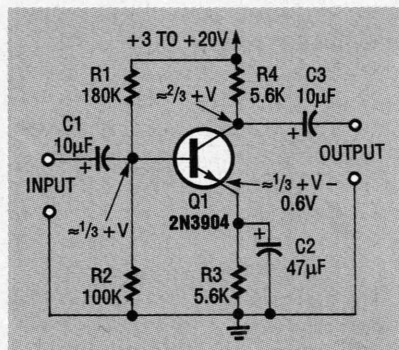


FIG. 13—COMMON-EMITTER amplifier with voltage-divider biasing.

part of a water-level alarm circuit. The relay can actuate an auxiliary alarm circuit if rising water shorts the probes.

Figure 9 is a schematic for another ultra-sensitive, dual-transistor relay driver (based on Fig. 5). An input of about 0.7 volt at 40 microamperes is needed to actuate the relay. The 22-kilohm resistor between the base and the emitter of Q1 ensures that both Q1 and Q2 are fully cut off if the input terminals are open-circuited.

Linear biasing circuits

Figure 10 shows how a common-emitter amplifier can become a linear AC amplifier. This is done by applying a DC bias current to the base of Q1 so that the collector's quiescent value is about half its supply voltage. This bias permits the amplifier to accommodate high AC output swings without distorting them. The AC-input signal is applied between the base and ground of Q1, and the AC-output signal is taken between the collector and ground.

To design an AC amplifier as shown in Fig. 10, first decide on the value of load resistor R2.

The lower this resistance value, the higher will be the amplifier's upper cut-off frequency. (This is due to the lower shunting effects of stray output capacitance on the effective impedance of the load.)

Moreover, a low value of R2 will cause a higher quiescent operating current in Q1. In Fig. 10, resistor R2 has a value of 5.6 kilohms, a compromise that will amplify a frequency of about 120 kHz and draw about a milliampere of quiescent current from a 12-volt supply. To bias the output to half the supply voltage, R1 must have a value of

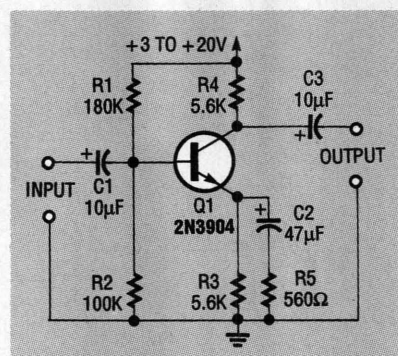


FIG. 14—COMMON-EMITTER amplifier with a fixed gain of 10.

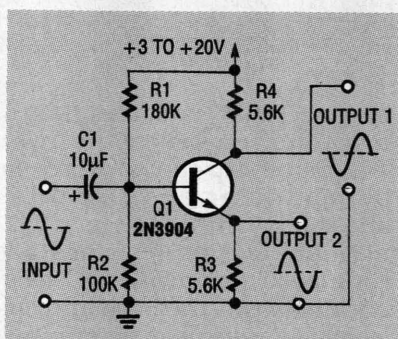


FIG. 15—PHASE-SPLITTER with unity gain.

$R2 \times 2h_{FE}$
 $R1 = 2 \times 200 \times 5600 \approx 2.2 \text{ megohms}$
 where R2 is 5600 ohms and h_{FE} is 200

Input impedance and voltage gain for this circuit are both determined by the impedance of transistor Q1's internal base-to-emitter junction. For a 2N3904, this value will be about $25/I_C$, where I_C is the collector current value in milliamperes. (This impedance is 25 ohms at 1 mA, 12.5 ohms at 2 mA, or 50 ohms at 0.5 mA.)

The input impedance into the transistor base is:

$$Z_{IN} = h_{FE} \times 25/I_C \\ = 200 \times 25 = \approx 5 \text{ kilohms at } 1 \text{ mA (shunted by } R1)$$

The voltage gain (A_v) of the circuit in Fig. 10 is calculated as follows:

$$A_v = R2/25/I_C = 5600/25 \\ = \approx 200 \text{ at } 1 \text{ mA}$$

This gain figure (which translates into approximately 46 dB, also determines the theoretical maximum attainable upper frequency response, measured at the -3-dB point of the frequency response curve. It equals f_T/A_v where f_T is the transit time limit of minority carriers across the base region. (f_T is the frequency at which h_{FE} decreases to unity, and is the measure of the transistor's high-frequency performance.) The f_T of the 2N3904 is about 300 MHz.

Therefore, the maximum frequency response of the circuit in Fig. 10 (ignoring the effects of stray capacitance) is $300 \text{ MHz}/200 = \approx 1.5 \text{ MHz}$.

A shortcoming of the Fig. 10 circuit is that its quiescent

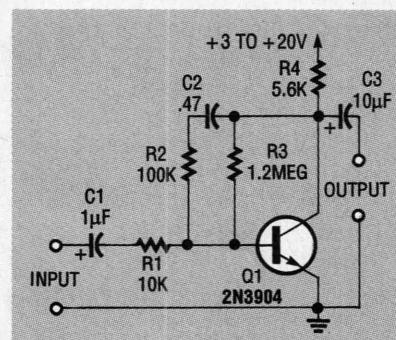


FIG. 16—ALTERNATIVE AMPLIFIER with a fixed gain of 10.

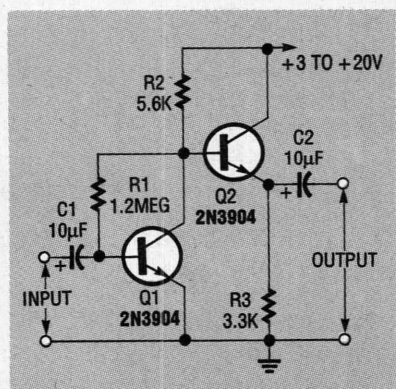


FIG. 17—WIDEBAND, TWO-STAGE common-emitter amplifier.

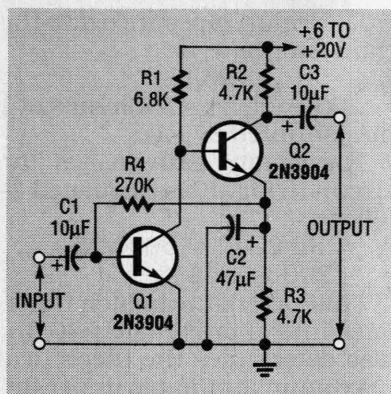


FIG. 18—HIGH-GAIN, TWO-STAGE amplifier.

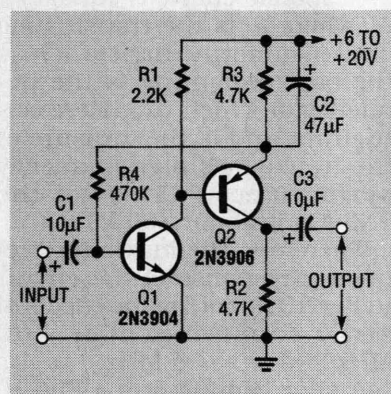


FIG. 19—ALTERNATIVE HIGH-GAIN two-stage amplifier.

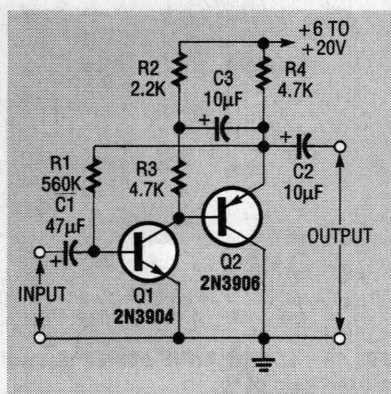


FIG. 20—BOOTSTRAPPED HIGH-GAIN two-stage amplifier.

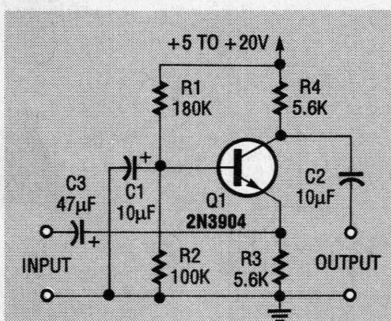


FIG. 21—COMMON-BASE AMPLIFIER with two output terminals.

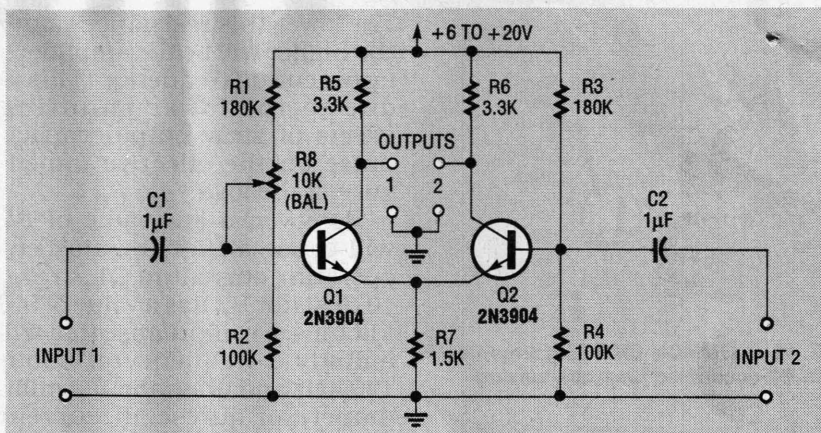


FIG. 22—DIFFERENTIAL AMPLIFIER OR "LONG-TAILED" pair.

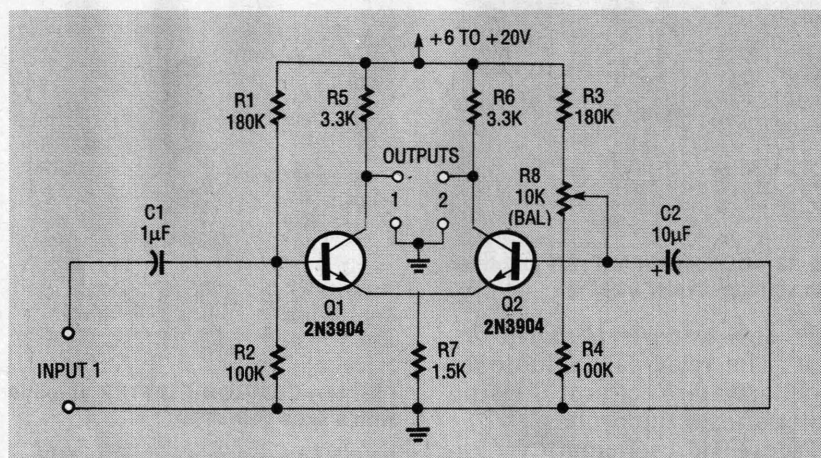


FIG. 23—PHASE-SPLITTER based on the differential amplifier.

biasing point is a function of transistor current gain (h_{FE}). This handicap can be overcome by modifying the circuit as shown in Fig. 11. Here 1.2-megohm biasing resistor R1 is connected to provide DC feedback between the base and collector. Its value was determined by multiplying R2 (5.6 kilohms) by h_{FE} (200).

Feedback occurs when any shift in the output biasing point takes place due to variations in h_{FE} , ambient temperature changes, or a shift in critical value of a connected passive component. Feedback automatically introduces a counteracting current in the base-current bias, which tends to cancel the original shift.

The Fig. 11 circuit offers the same bandwidth and voltage gain as the circuit in Fig. 10, but it has a lower input impedance because the AC feedback of

the Fig. 11 circuit reduces the effective value of resistor R1. That resistor, which shunts the 5-kilohm base impedance of Q1, is reduced by the voltage gain of 200. As a result, total input impedance is 2.7 kilohms.

The shunting effects of the biasing network can be eliminated with two feedback resistors and AC decoupling capacitors, as shown in Fig. 12.

The ultimate in biasing stability can be obtained with *potential-divider* biasing as shown in Fig. 13. Here, the voltage divider formed by junction of resistors R1 and R2 applies a quiescent voltage slightly higher than one-third of the supply voltage on the base of Q1. "Voltage-follower" action reduces the supply voltage at Q1's emitter by 0.6 volt.

As a result, one-third of the supply voltage is developed

Continued on page 86

BUILD A DIVERSITY ANTENNA AND IMPROVE THE PERFORMANCE OF ANY CAR STEREO!

WHEN LISTENING TO AN FM STATION in your car, have you ever noticed the sudden onset of noise—pops, clicks and hum—that lasts for just a short time? Maybe you stopped at a traffic light, heard the interference, but found that it disappeared when you drove away—less than a car length.

This audio annoyance could be caused by local sources of noise, or it could be caused by *multipath*—the convergence of FM signals at your car's antenna that arrived by taking different paths from the FM transmitting antenna. The interference is commonly called "picket fencing" because it comes and goes as you drive, much as your view changes as you walk by a picket fence.

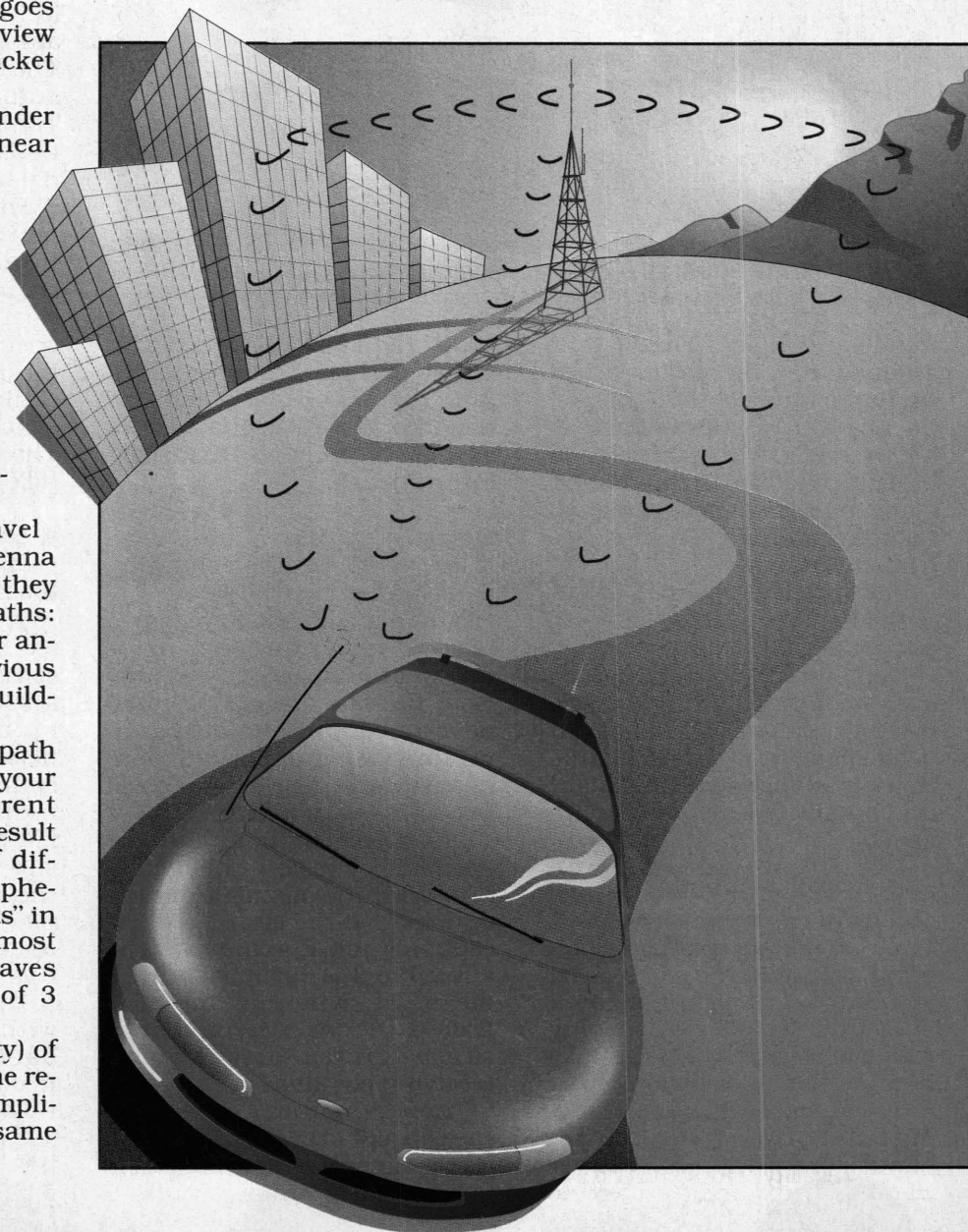
You could be stopped under overhead power lines or near neon lights, motors or relays that could introduce noise into the FM receivers of even the latest model, high-priced cars. However, if you look around and find no obvious sources of electrical noise, consider the possibility that the interference is caused by multipath.

As FM radio waves travel from the transmitting antenna to your receiving antenna, they can take many different paths: Some travel directly to your antenna; others take more devious routes as they bounce off buildings, hills or mountains.

Figure 1 illustrates multipath radio waves converging on your car's antenna with different phase relationships as a result of traveling over paths of different lengths. This same phenomenon can cause "ghosts" in television reception. It is most noticeable with radio waves that have wavelengths of 3 meters or less.

The strength (and quality) of the received FM signal is the resultant of the phase and amplitude of all the waves at the same

Eliminate multipath noise from your FM radio as you drive with this diversity circuit that switches antennas.



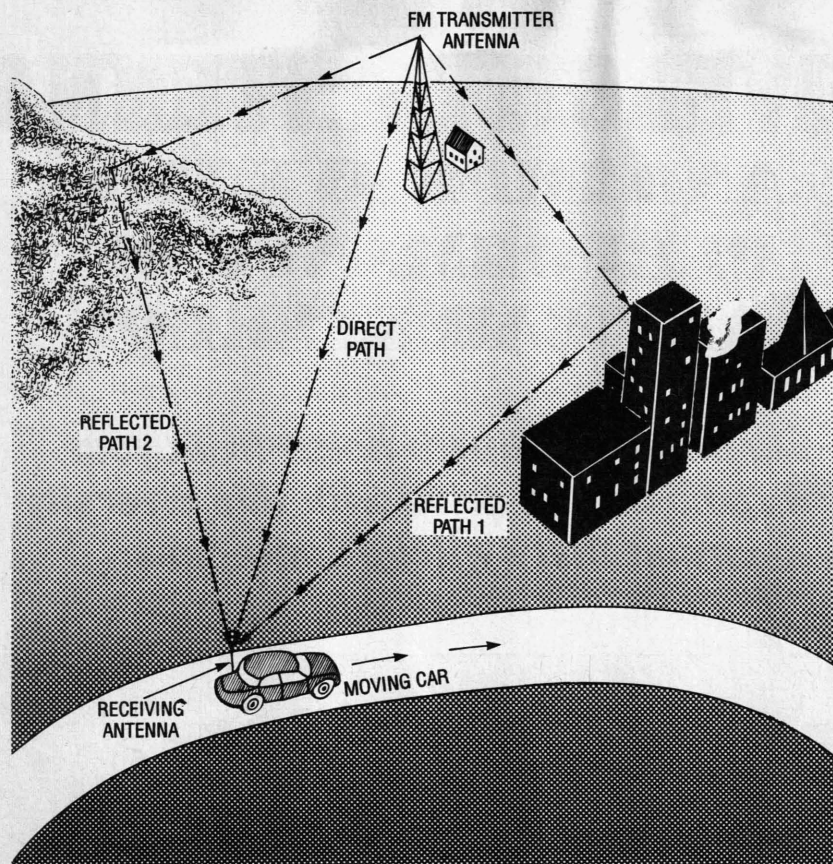


FIG. 1—MULTIPATH INTERFERENCE IN FM RECEPTION is caused by signals arriving out-of-phase at the antenna after traveling over paths of different length.

frequency which arrive at your antenna from the station tuned in. Radio waves of the same frequency that are out-of-phase, as shown in Fig. 2, can cancel each other in certain locations and blank out the received signal, regardless of the FM station's transmitter power.

However, the effects of multipath are more likely to show up as partial cancellation of the received signal accompanied by extraneous noise. If you keep driving, you will soon pass out of this "noisy" region. Fluctuations in signal strength might occur just a quarter wavelength apart.

The FM broadcast band covers the radio-frequency spectrum from 88 to 108 MHz. Thus at the approximate mid-band frequency of 100 kHz, wavelength is 3 meters or about 3¼ yards. That's why moving your car only a few feet can take it out of the noisy region.

By contrast, the amplitude-modulated (AM) broadcast band, covers the much lower frequency spectrum of 540 to

1600 kHz. Thus at 1000 kHz, an AM signal has a wavelength of 300 meters—100 times the length of the FM signal. That's why AM reception is unaffected by multipath.

Diversity systems

In the FM reception situation described, if instead of moving the car out of the noisy region, a second antenna were positioned at least 30 inches away from the first, reception could be restored. Unfortunately, connecting two antennas simultaneously to a single car radio will not solve the problem.

The signals from the antenna in the noisy region would combine with the signals from the antenna "in-the-clear," and reception would not improve. The answer to this dilemma is find a means of switching automatically to the one of two antennas situated in the most favorable receiving position.

There is nothing new about the concept of switching antennas to improve reception. One method called *diversity recep-*

tion was developed in the early days of radio to counter the effects of "fading" in shortwave reception. Shortwave or high-frequency (HF) signals, are capable of traveling thousands of miles by "bouncing" off ionized layers 100 kilometers or higher in ionosphere. They were once the best method for long-range communication, and fading could break that communications link.

In those early high-frequency diversity systems, two separate antennas positioned several miles apart fed two separate receiver sections. Electronic circuits compared the relative strengths of the two received signals, and automatically selected the strongest for further amplification and reception. The selection was performed by automatic gain control (AGC). The output DC level was proportional to the strength of the signal being received.

The same circuitry could improve mobile FM reception, but two complete receivers would be required—obviously impractical and expensive. Moreover, opening a standard automotive receiver case to add circuitry could pose a problem due to space and power limitations. The circuit described in this article solves that problem.

A novel FM system

In stereo FM broadcasting, the transmitter encodes left and right channels as sum and difference signals. The difference signal, L - R, modulates the 38-kilohertz sub-carrier necessary for decoding the stereo channels at the receiver. It is not transmitted because of bandwidth restrictions. Instead, the FM station transmits a *pilot subcarrier* at 19 kilohertz, half the subcarrier frequency. This pilot phase locks a 38-kilohertz oscillator in the receiver to decode the stereo signal.

The 19-kHz pilot subcarrier is within the audio bandwidth, but its amplitude is so low that it doesn't disturb the listener. However, the presence of this pilot subcarrier makes possible the FM diversity reception system discussed in this article.

The FM diversity circuit monitors the 19-kHz pilot sub-carrier signal.

Steadiness of the reception of this 19-kHz pilot signal in the audio portion of the FM transmission is an indication of the quality of the received signal. Whenever this signal falls below a specified threshold value, it will be lost in background noise. The threshold establishes the criterion for switching antennas. In effect, the pilot threshold level functions in FM diversity as the AGC level functions in HF diversity.

How FM diversity works

A second antenna, installed on your vehicle as far away from the original equipment antenna as practical, provides the second FM signal. Figure 3 is a simplified block diagram of the diversity system.

The cables from both antennas are connected to the electronic antenna switch. The 19-kHz pilot signal from the receiver's audio output is passed through a high-gain bandpass active filter which attenuates audio programming that is much stronger than the pilot signal. After amplification, the pilot subcarrier becomes the reference frequency for a phase-locked loop (PLL) circuit. The output of the PLL locks to the 19-kHz pilot signal and functions as a subcarrier detector.

When the reference frequency becomes noisy, the PLL will lose "lock" and trigger the flip-flop whose output switches the state of the electronic antenna switch. This action switches the alternate antenna into the system while disabling the original antenna.

If that second antenna is positioned for better reception, the received signal will clear, and the PLL will again lock to the subcarrier and hold the switch in that state until the pilot signal drops out again. If the second antenna does not restore the pilot signal reception after a 0.1 second delay, the primary antenna is switched back on.

When the radio is receiving AM, the absence of a 19-kHz subcarrier will also reactivate the primary antenna that is tuned to the receiver for the best AM reception.

FM diversity circuit

Refer to schematic diagram Fig. 4. The audio signal from the FM receiver appears at connector J4. The two capacitors C11 and C16 in the audio input section bypass any DC components in the radio output or overvoltages that could be caused by miswiring. Trimmer potentiometer R22 controls the input level. The LF347 quad operational amplifier IC3 is an active filter with a gain of 50 at 19 KHz. It has four sections: a, b, c,

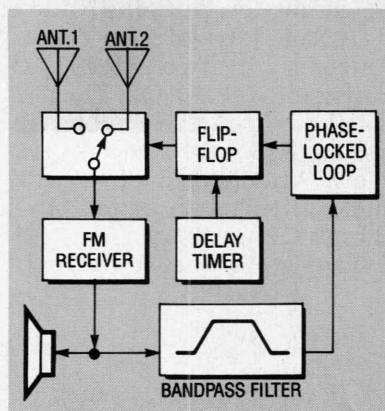


FIG. 3—BLOCK DIAGRAM OF THE FM DIVERSITY SYSTEM to overcome multipath interference or cancellation.

and d.

The active filter attenuates the audio so that the LM1800N phase-locked loop (PLL) IC1 can lock onto the 19-kHz pilot sub-carrier. With 2 millivolts input, the output level at pin 14 of IC3-d is about 100 millivolts. Light-emitting diode LED3 is the level indicator for IC3-d. Each of the four 47,000-ohm feedback resistors, R1, R6, R7, and R8, around the op-amp sections in IC3 has a 1% tolerance. The feedback capacitors (reading from left to right) C6, C4, C19 and C23, and the input capacitors C7, C5, C20, and C21 have closer 10% tolerances to assure that the filter will tune in the 19-kHz region.

Trimmer potentiometer R21 (in series with resistor R10 at frequency pin 15 of IC1) sets the PLL's operating frequency to 19 KHz. Resistor R9 and capacitors C2 and C12 form the loop filter between pins 13 and 14 of IC1 to set the PLL's locking characteristics including *capture time* and *capture range*. The values shown result in a 1-millisecond capture time and 2-kHz bandwidth. Bandwidth is not critical in this circuit because the center frequency is always 19 kHz. However, the wider the bandwidth, the faster the capture.

Every time the PLL locks to the incoming signal, it produces a low-level logic output at LAMP pin 7 of IC1. When the input signal is lost, pin 7 of IC1 goes high, toggling 74C74 CMOS dual flip-flop IC2-a. Com-

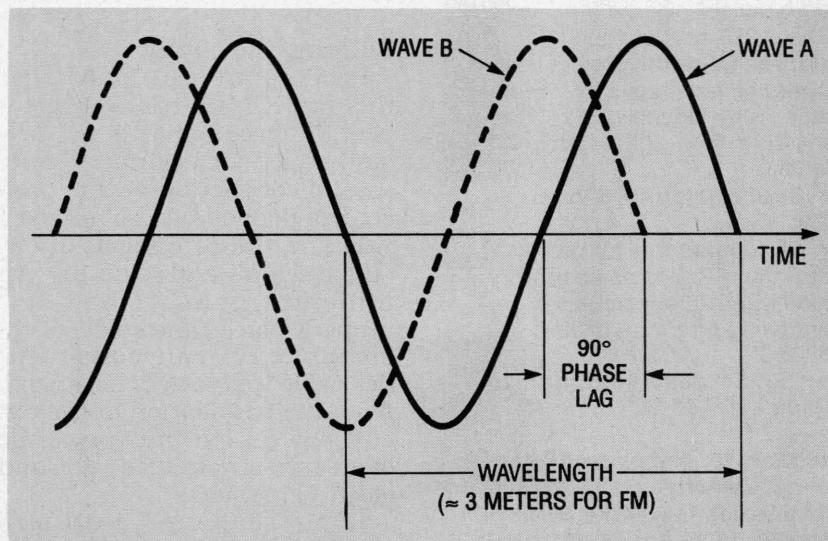


FIG. 2—OUT-OF-PHASE SINE WAVES represent out-of-phase FM signals arriving at a receiving antenna. They can combine constructively or destructively. An FM signal is about 3 meters long.

plementary output pins 5 and 6 of IC2-a control the ON/OFF states of the two MR901 RF transistors Q1 and Q2. They are switched through 33-kilohm resistors R13 and R12, respectively. When Pin 5 of IC2-a is high, Q1 turns on; when pin 6 is high Q2 turns on.

Transistors Q1 and Q2 are the

active components in two identical broadband, untuned, common-emitter amplifiers. Each has a gain of about unity in the AM broadcast band and about 6 dB in the FM band. That 6-dB gain overcomes cable and connector losses. The low AM gain prevents AM signal overload.

Transistor Q1's base is con-

nected to antenna input jack J3, for Antenna 1 and transistor Q2's base is connected to input jack J3 from antenna 2. The transistor collectors are connected, but only one transistor can be turned on at a time. As a result, the circuit works as a fast electronic single-pole, double-throw antenna switch.

The load inductance for Q1 and Q2 presents a high impedance at FM frequencies for good amplification. However, at AM frequencies this impedance is low, resulting in low amplification. The output signals at Q1 and Q2 feed back to the working antenna input jack.

The second half of the CMOS dual flip-flop IC2, section b, is a 0.1-second timer. If the FM pilot subcarrier is absent for more than 0.1 second, 0.1 μ F capacitor C17, charging through 1-megohm resistor R15, toggles flip-flop IC2-b, forcing pin 8 low. That low output presets the IC2-a flip-flop, biasing Q1 on and activating antenna 1. This feature is necessary for AM operation because antenna 1 is the car's original equipment or primary AM antenna.

Flip-flop IC2-a also forces Q1 on for non-stereo FM signals. The illumination of PILOT LED 3 indicates that the PLL is locked to the pilot signal. The illumination of ANT 1 LED2 indicates that antenna 1 is active, and the illumination of ANT 2 LED1 indicates that antenna 2 is active.

Building the circuit

Readers are cautioned that this project is a relatively complex RF circuit that is not recommended for beginners. Successful completion of this project will depend on the builder's skill and the care taken in circuit assembly and soldering. An understanding of how improperly placed and soldered RF components can cause undesirable feedback is necessary. Also, the installation of the circuit calls for current knowledge of modern automotive radio and electrical systems.

Even seasoned builders should pay particular attention to details and work cautiously, especially in the RF adjustment,

PARTS LIST

All resistors are 1/4-watt, 5%, unless otherwise specified.

R1, R6, R7, R8—47,000 ohms, 1%
R2, R11, R23, R27, R29—10,000 ohms
R3, R4, R5, R18, R24, R28, R32—1800 ohms
R9—3,300 ohms
R10—13,000 ohms
R12, R13, R25, R26—33,000
R14—100 ohms
R15—1 megohm
R16, R19—10 ohms
R17—1000 ohms
R20—680 ohms
R21, R22—20,000 ohms trimmer potentiometer, PC mount, top adjust, Spectrol 321 or equivalent
R30, R31—150 ohms
R33—100,000 ohms

All capacitors are 20%, 25-volt, unless otherwise specified

C1, C2—0.033 μ F
C3—390 pF
C4, C5, C6, C7, C19, C20, C21, C23—0.001 μ F, 10% polyester film.
C8—0.068 μ F
C9, C10, C13—470 pF ceramic disc
C11, C12, C15—0.01 μ F
C14, C18—0.001 μ F ceramic disc
C16—1.0 μ F
C17—0.1 μ F
C22—680 pF
C24—10 μ F solid tantalum electrolytic
C25—47 μ F solid tantalum electrolytic

Semiconductors

IC1—LM1800N phase-locked loop, National Semiconductor or equiv.
IC2—74C74 CMOS B dual flip-flop, National Semiconductor or equivalent
IC3—LF347 BiFET quad operational amplifier, Motorola Semiconductor or equivalent
Q1, Q2—MR901 or MR911 RF NPN transistor, Motorola or equivalent.
D1—IN914 silicon diode
LED1, LED2, LED3—light-emitting diodes, T-1 package, red, green, yellow

Other components

L1—22 μ H RF choke
J1—power jack—Radio Shack 274-1565 or equivalent
J2—phono jack, RCA style
J4, J5, J6—Antenna connectors
Miscellaneous: automotive antenna, Motorola type; cigarette-lighter power cord (Radio Shack 270-1533-C or equiv.); plastic case (Serpac No. 132 from Serco); extension cable, male-to-male (Radio Shack 12-1314 or equiv.); extension cable, male-to-female, 12-foot, (Radio Shack 12-1311 or equiv.); audio cable, RCA phono-type; one 14-pin DIP socket; two 16-pin DIP sockets; hook-up wire, stranded, insulated 24 or 26 AWG; plastic clamp (Radio Shack 278-1641 or equiv.); solder.

Optional miscellaneous: adhesive-mount automotive adhesive antenna (Radio Shack 12-1325 or equiv.); automotive antenna (No. 03CH7516N, J.C. Whitney, P.O. Box 8410, Chicago IL 60680 (312) 431-6102); miniature power plug (Radio Shack 274-1567 or equiv.); fuse holder, in-line (Radio Shack 270-1213 or equiv.).

Note: The following project components and parts are available from Videoart, 4123 Middlefield Rd., Palo Alto, CA 94303:

- Double-sided PC board—\$28.00
- All components for completing the PC board assembly plus three antenna connectors but excluding the PC board—\$23.00
- Plastic case specified in Parts List—\$7.50

Add \$6.50 for shipping and handling. California residents add local sales tax. Make all payments as either in certified check or money order. Please allow 4 to 6 weeks for delivery.

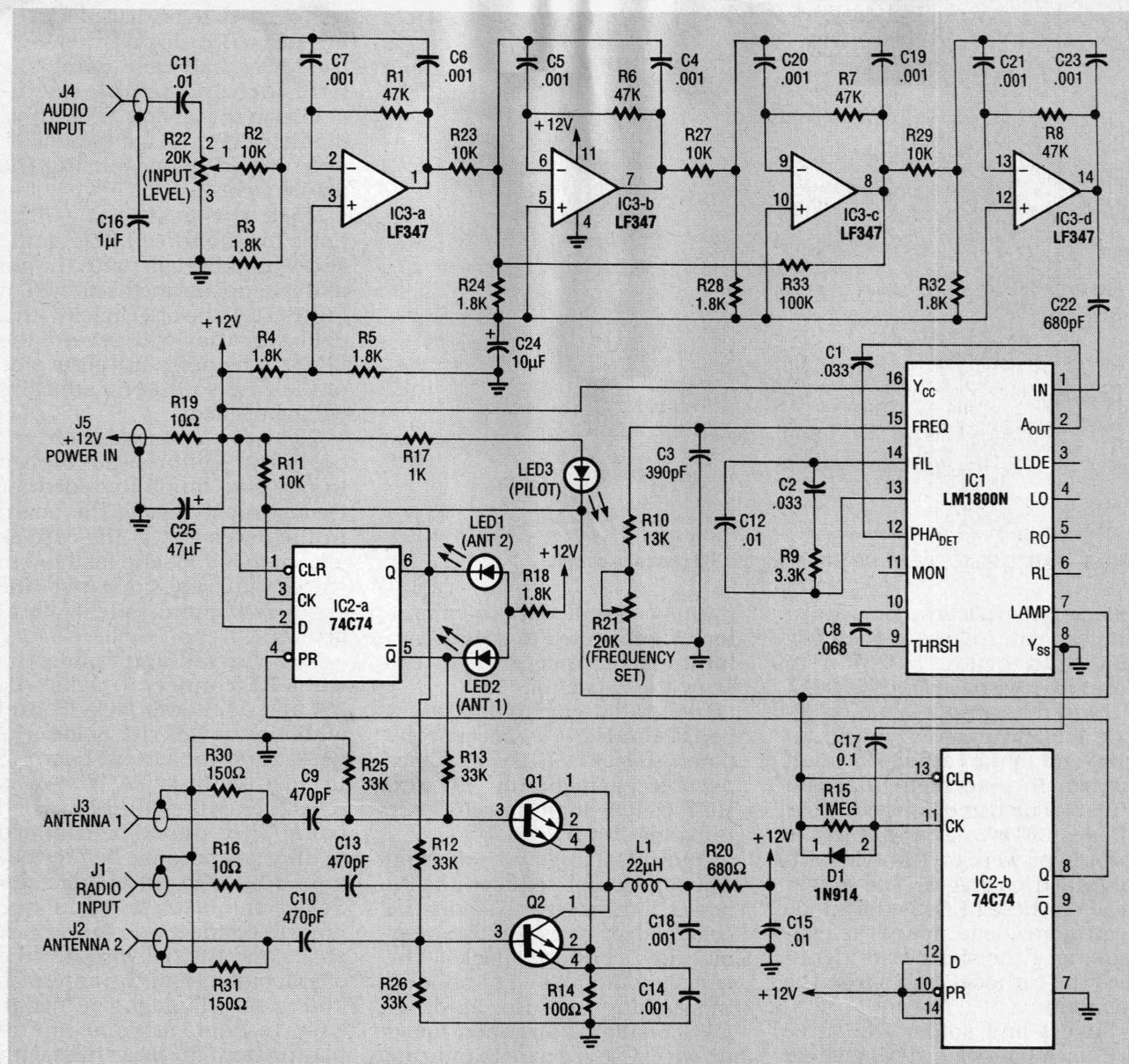


FIG. 4—SCHEMATIC DIAGRAM for the FM diversity circuit.

test and installation phases of the project. Experience in working on automotive electrical and entertainment systems will be helpful.

Because the circuit must receive clear high-frequency FM signals, PC board construction is recommended. A partial kit, including an etched and drilled double-sided PC board with a ground plane, is available from the source given in the Parts List. Foil patterns are provided for those who want to make their own boards.

The ground plane is copper foil laminated over most of the component side of the board to

ensure stable RF reception. It shields the active components to prevent inadvertent signal radiation, thus preventing unwanted oscillations. All of the electronic components are standard parts, stock items from mail-order distributors and most electronics retail stores.

All wiring must be as short as practical. For example, the leads of collector inductor L1 must be short to prevent unwanted oscillations. However, parts placement and wiring in the audio section is not critical.

Refer to both the schematic Fig. 4 and the parts placement diagram Fig. 5. Follow accepted

parts placement practices, and do all soldering with a fine-tipped soldering pencil rated 30 watts or less, preferably with a temperature control set to 700°F. The cleanliness of the PC board and component leads is important for quality soldering. Be sure that all solder joints are smooth and shiny; cold solder joints are usually dull gray and irregular.

Use sockets for the all ICs, and observe the location of pin 1 when installing all sockets. Mount the eight axial-leaded 0.001 μ F polyester capacitors C4 to C7, C19 to 21, and C23, the 390 pF C3 and the 0.047 μ F C8 vertically. With needle-nose

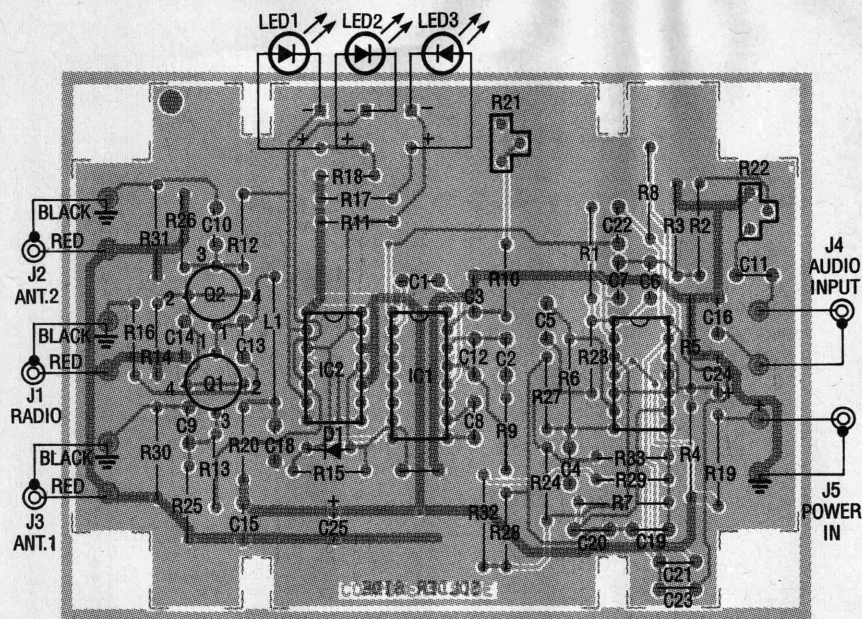


FIG. 5—PARTS PLACEMENT DIAGRAM for the FM diversity circuit.

pliers, grasp a lead at one end of the capacitor close to the body and bend the lead back 180° to form closely spaced radial leads.

Find the cathode band on diode D1 and insert the lead on that end in the cathode location shown in placement diagram Fig. 5. The three light-emitting diodes LEDs 1, 2, and 3 in the prototype were in T-1-style radial-leaded packages. The longer lead on these LEDs is the anode lead (arrowhead side of the symbol), and the short lead identifies the cathode lead (bar in the symbol).

Insert and solder the three LEDs without cutting their leads. Mount them so they stand off the board about 1¼-inches so the leads can be bent to insert the reflectors into the pre-drilled holes in the case after the completed board is mounted to the lower case half.

Transistors Q1 and Q2 have flat leads to minimize inductance. Position the transistors carefully on the board before soldering them. *Caution:* The transistors' orientation and spacing with respect to each other and the other RF components is critical.

Install the two polarized capacitors, tantalum dipped, radial capacitor C24 and aluminum electrolytic C25, according to the polarity marks shown on Fig. 5. *Caution:* Be sure the

minus (–) side of both capacitors is connected to ground; an improper connection can destroy the capacitor.

Integrated circuit IC2 must be a CMOS B 74C74 because the power source is +12 volts. Comparable parts in the HC and HCT CMOS logic families are rated for only 5 volts.

After inserting and soldering all of the components on the PC board, cut, insert and solder the color-coded wires for the case-mounted connector jacks. The wires should all be cut about 3 inches long from No. 22 or 24 AWG insulated stranded hook-up wire. Use red wire to indicate positive (+) and black to indicate negative (–) ground for power jack J5. Use any other color for the audio input signal to jack J4, but use black for the ground connection there also.

Insert and solder lengths of twisted black and red wire for the RF connector jacks J1, J2 and J3 with the red wires for the antenna and radio input signals and black wires for ground.

Packaging the circuit

The plastic enclosure specified in the Parts List is recommended because the circuit board described in this article is sized to fit snugly in it. Refer to Fig. 6 for the proper orientation of jacks with respect to the circuit board, and mark the cen-

ters of holes to be drilled for the five jacks J1 to J5 on the ends of the lower half case using the hole-forming templates included in this article. Exact hole diameter and shape should be determined by measuring the actual jacks.

Mark the positions of the holes for the three LEDs in the side wall of the case with the aid of the template and as shown in Fig. 6. Drill the LED holes with a drill size that will permit the LEDs to be press-fit in the side of the case so that no adhesive will be needed.

Clean the completed circuit board with cotton swabs dipped in cleaning fluids intended for that purpose. Insert the board in the lower half of the case as shown in Fig. 6, and fasten it to the two internal sidebars of the case with small sheet-metal screws.

Next, install and fasten the three RF connector jacks, J1, J2, and J3, power jack J5, and audio input jack J6. Solder the wires from the circuit board to the lugs on the jacks. Wire jacks J1, J2, and J3 with the twisted pairs. They should be trimmed as short as possible before they are soldered to the connectors. Solder the black wires for the ground connections to the jack shells and the red wires for the signal paths from the antennas and radio. (It might be necessary to remove some metal plating from the jack shells with emery cloth or a file to obtain a secure solder joint.)

Carefully bend both LED leads together so that the reflector body can be press-fit through the previously drilled holes on the side of the case.

Double check the completed assembly carefully. A magnifying glass will be helpful. Check for any incorrectly inserted components, solder bridges, or cold solder joints.

Identify all the jacks and LEDs on the outside of the case with a waterproof pen or dry transfer lettering. Label LED3 as PILOT, LED1 as ANT 2 and LED2 as ANT 1. Label J1 as RADIO INPUT, J2 as ANT 1 and J3 as ANT 2. Label J4 as AUDIO INPUT and J5 as "+12 V." Cover the

labels with transparent tape or coat them with nail lacquer. Figure 8 is a photo of the author's assembled prototype.

Testing and tuning

Before applying power to the board, measure the resistance between the positive power supply connection and ground. It should be 3000 ohms or higher, after the filter capacitor charges. If it is lower, recheck the circuit for shorts or incorrectly installed ICs.

The power source required to perform these tests can be a 12-volt nickel-cadmium battery, a 12-volt lead-acid battery, or a 12-volt DC wall-outlet adapter. If you use an adapter, be sure it has a standard 2.1 millimeter diameter axial pin in the plug. Read the label on the adapter to be sure that the positive (+) conductor of the plug is the axial lead and the negative (-) conductor is the shell.

Put a ¼-ampere fuse in series with the power supply to pre-

vent damage to the circuit if there are undetected shorts. An FM radio or tuner with an earphone or speaker plug connector is also required. The test setup shown in Fig. 7 emulates the wiring circuitry of an automobile installation.

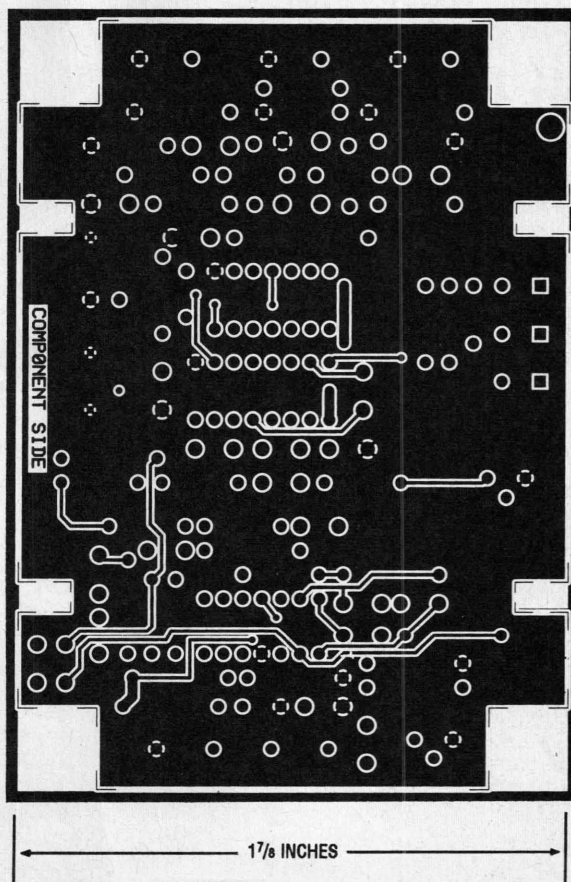
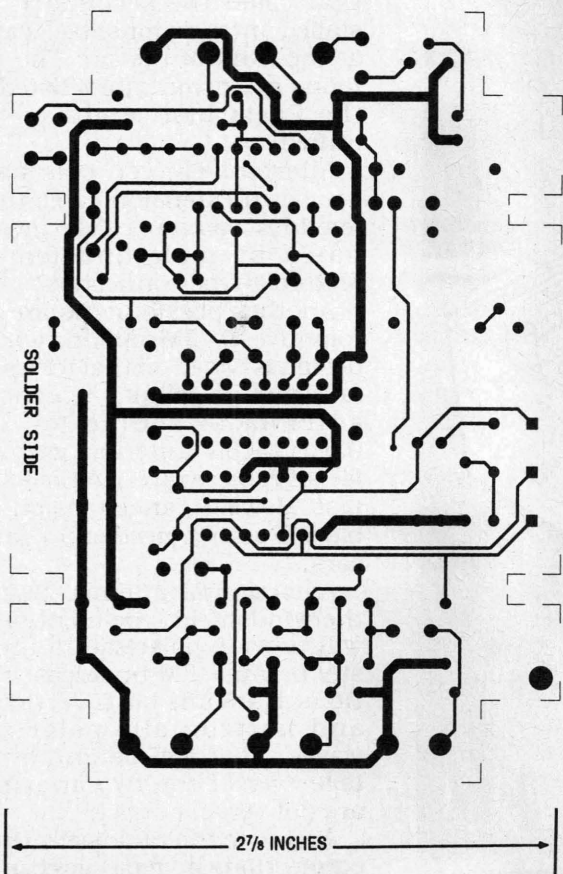
Tune the FM receiver to a stereo FM station. (The stereo indicator should be illuminated). Set the volume control from one quarter to one third of its maximum angle. Then connect the audio input of the diversity circuit to the FM radio's head-phone or speaker.

Apply power to the system. With a plastic alignment tool or small screwdriver, adjust audio input-level potentiometer R22 to mid-position. Then adjust PLL frequency-set potentiometer R21 slowly until LED3 is illuminated. After this adjustment, LED3 will track the radio's stereo indicator to verify the presence of the stereo pilot subcarrier signal with different thresholds.

The radio indicator works with all stereo stations, and the diversity circuit depends on the audio volume and the setting of input-level potentiometer R22. Either LED1 or LED2 should be on. Turn down the volume control on the FM radio until LED 3 extinguishes. Observe the LED1 and LED2 pair: They should alternate between on and off each time LED3 goes out.

Turn the volume-control knob up and down to check the switching action. Turn the volume down once or twice so LED2 stays on, then keep the volume low. After 0.1 second, LED1 will light and LED2 will go off. That indicates the one-second timer is working.

Connect the RADIO INPUT cable from J1 to the FM radio's antenna input jack. If the existing FM radio receiver does not have external antenna connections, retract its antenna completely, and convert it to the input connection with wires connected by



FOIL PATTERN FOR THE COMPONENT SIDE OF THE PC BOARD. FOIL PATTERN FOR THE SOLDER SIDE OF THE PC BOARD.

two alligator clips. The second connection is for ground, the outer part of the cable or jack.

Unless this connection is made with a coaxial cable, the antenna will not function correctly, and the test will be invalid. Light-emitting diode LED2 will be illuminated after the one-second timeout. Antenna 1 will be the active antenna. Conversely, LED1 will indicate that antenna 2 is active.

Connect two test antenna cables to their jacks in the diversity circuit. One of these could be the extra antenna you purchased for installation in your car. A three-foot length of insulated hookup wire stretched vertically will serve as the second antenna. Regardless of what you use as an antenna, it must make contact with the axial conductor of the connector. A banana plug can serve as a makeshift connector.

Tune in an FM station, check

to see which LED is illuminated, and then disconnect the related antenna. This step will permit the circuit to switch to the alternative antenna, as shown by LED illumination. If the signal from the FM station selected is strong, switching action can be prevented if the antenna connections are not well shielded.

Choose a different station and repeat the test. Then repeat the test for the other antenna. Adjust the volume control to command antenna switching. When the desired LED turns on, remove the corresponding antenna, and the ability of the LEDs to turn on and off should be restored.

This test does not reproduce actual circuit operation because it is not possible to simulate, at a fixed location, the FM radio reception conditions to which your vehicle is exposed. In a car installation, the proper

connectors and coaxial cables form well shielded connections without RF leakage. Nevertheless, this bench test can demonstrate that the RF switching is functional.

If you can perform this test with an actual auto radio receiver and the coaxial cables recommended in the Parts List, the test will still be a realistic approximation of an automotive installation.

Installation in a car

Refer again to Fig. 7, and install the second antenna on your car. The greater the separation between the primary and secondary antennas, the more effective will be the diversity circuit's operation. If, for any reason, you do not want to install a second full-size antenna, you can purchase a flat antenna that adheres to the windshield glass.

The cable from the originally installed antenna probably will not reach the diversity circuit unless the antenna was installed at the rear of the car. In that case, the secondary installed antenna must be located at the front of the car. The antenna extension cables listed in the Parts List might be required.

Alternatively, you can make your own extension cable from low-loss coaxial cable terminated by male and female Motorola-type connectors.

Another possibility is an automotive AM/FM antenna whose design is based on that for a surface-mount cellular telephone antenna. A source for the 03CH7516N antenna with 17 feet of cable is given in the Parts List. However, an extension cable will not be needed for most cars.

A horizontal antenna such as the windshield type mentioned will provide polarization diversity because FM broadcast stations transmit both vertically and horizontally polarized waves. This will be an advantage even if the two antennas are not very far apart.

Access to the radio's electrical connections in most cars can be gained by removing the front

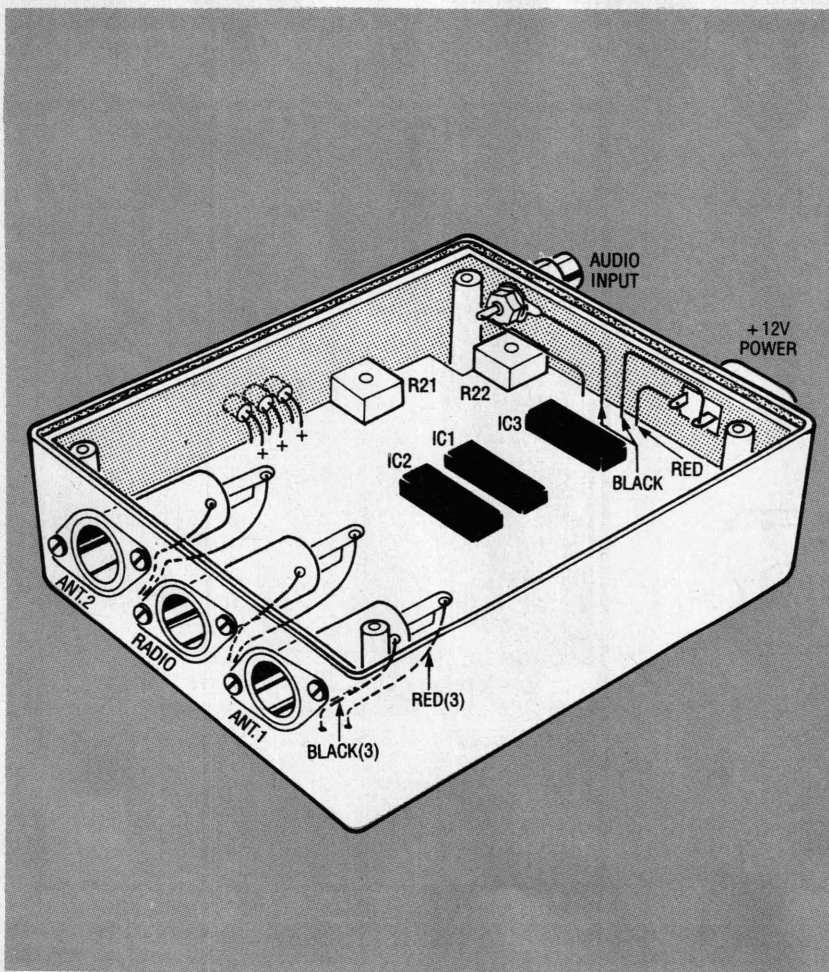


FIG. 6—ASSEMBLY AND WIRING OF CASE-MOUNTED components.

panel of the radio, and pulling it out to expose the antenna and speaker connections. Consult the maintenance manual for your car radio for details on how to remove the radio without damaging it.

The latest model car radios have RCA audio output connectors. Those will make it easy to provide the audio for the diversity circuit.

Alternatively, the audio signal source will be the radio's speaker output terminals. Any of the stereo outputs will provide the signal depending, of course, on the stereo BALANCE control setting. Identify all of the speakers' terminals and wiring. Consult your user's manual or read the labels on the wires.

Radio manufacturers do not all follow a uniform wiring color code, so the functions of the wires cannot be determined reliably from their colors. However, several leading manufacturers have agreed on green and gray for the left and right speakers, respectively, and black for common or ground. The ground (or low side) goes to the ground side of the audio input jack of the diversity circuit. Only one audio source is required for the operation of this circuit.

Caution: Do not connect the unit to the output of an external power booster.

Unfortunately, you have no guarantee that your radio's power amplifier will have sufficient bandwidth to allow the 19-kHz subcarrier to pass. Tap the audio from the input cable to the power amplifier. If resistive "faders" have been installed, tap the audio signal upstream of them.

Any connections spliced to the speaker wires must be insulated with quality electric tape so they will not be shorted to ground. Connect the original equipment car antenna to the antenna input jack ANT 1. Then connect the second antenna to input jack ANT 2. Connect the diversity circuit to the FM receiver with an extension cable terminated by two male plugs, one end plugged into the RADIO INPUT jack.

The required 12-volt power

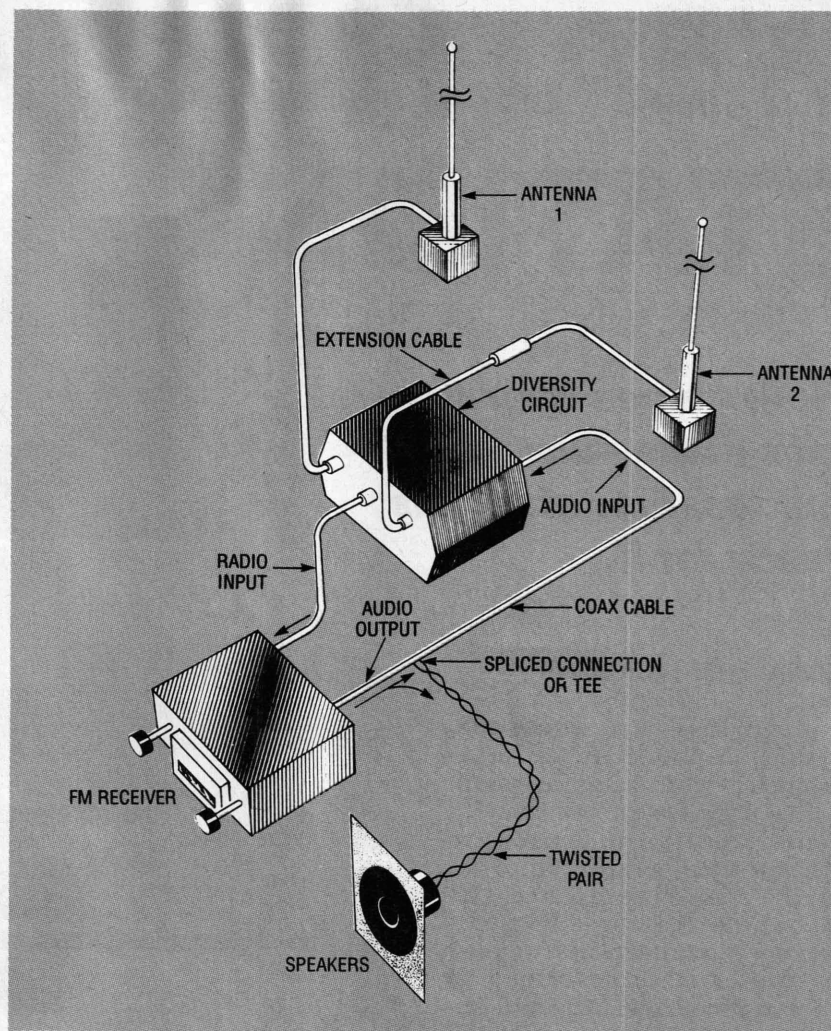


FIG. 7—THE FM DIVERSITY SYSTEM installed in an automobile with one antenna, a radio and speakers.

can be obtained from a fused cigarette lighter adapter cable. Install a $\frac{1}{4}$ - or $\frac{1}{8}$ -ampere fuse in the fuse holder to protect the diversity circuit. However, if you want a more permanent connection, you can make one with an in-line fused cable for the power connection to the spare lugs usually available in the car's fuse box. See the Parts List.

Turn on the receiver and plug in the adapter. Tune to a stereo station, and set the receiver for normal listening while driving. Readjust input-level trimmer potentiometer R22 for this normal audio volume setting in the car until LED3 remains on without blinking. Set the tone or treble control to at least one-quarter of a turn towards maximum. If this is not done, the 19 kHz pilot subcarrier signal will

be too low. If LED3 doesn't light up, readjust the PLL frequency-set trimmer potentiometer R21 until it does.

Verify that LED1 and LED2 change state each time you lower the volume. Then change stations. That also will control the LED1 to LED2 switching. Input-level trimmer potentiometer R22 controls switching sensitivity. If it is set too low, LED3 will not light. Set it so LED3 remains on without blinking when a clear stereo station is being received. Trimmer R21 must be set in the middle of its "lock" range, the span between the two points in wiper rotation where LED3 turns on and off.

Check the 19-kHz pilot subcarrier level at the audio output of a receiver with a high-Q band-

continued on page 85

consumer is happy with his Hitachi rice cooker, vacuum cleaner, or washing machine, he is likely to seriously consider a Hitachi audio component. In other words, the brand names of audio-equipment manufacturers are a familiar part of Japanese life. Large Japanese industrial companies also encourage their employees with various savings plans and discount arrangements to invest in their "house brand" audio equipment. The result of all this has been the development of a large number of hard-core audiophiles plus a higher level of interest on the part of the general public.

CD costs

Unlike the situation with conventional record players where, by and large, the more you paid the better they got, even the cheapest CD players provide excellent quality. Player costs have probably come down about as low as they are going to get. I've seen units advertised for slightly over \$100.

I still don't understand the economics of player pricing. They are not like memory ICs whose cost per unit falls radically with time and sales because the inherent "parts" cost is insignificant to start with. CD players are crammed with precision-made mechanical parts that, it seems to me, are going to continue to keep the prices from going any lower. I'm grateful—and so should the rest of you be—that the prices of CD players have fallen as far down as they have.

Disc prices, on the other hand, have not fallen significantly from the 1983 initial introductions at \$15 to \$20. Interestingly, during a 1981 report on a visit to Denon's Tokyo recording studio, I quoted an engineer who said that, unlike the situation with CD players, the cost of the disc itself would not come down substantially with time. However, considering the effects of inflation and the significant rise in the yen with respect to the dollar, CD's are now something of a bargain. At least, if sales figures are an accurate indication, the music listening public seems to think so.

In any case, Happy Tenth Birthday to the Digital Compact Disc—and long may it rotate. Ω

DIVERSITY ANTENNA

continued from page 39

pass filter and an oscilloscope. Five millivolts of the pilot sub-carrier will ensure proper system operation.

Temporarily fasten the diversity circuit to the automobile dashboard with duct tape, but put a cardboard hood over the LED lamps so that they can be seen more easily in daylight. Take the system for a test drive. Ask a passenger to check the switching action by observing the three LEDs as you drive.

When you are satisfied that the diversity circuit works, mount it in under the dashboard where it cannot be seen by passengers in the front seat, but where you can see the illuminated LEDs by ducking your head down close to the seat. Because of the wide variations in the size and shape of the space available under auto-

mobile dashboards and the limited space available for mounting an electronics case, universal mounting directions cannot be given. After selecting an appropriate location where the input and output wires and cables are free of stress, you can clamp the package in position with an adhesive-backed plastic clamp, a strap that you fashion from plastic or light metal, or even duct tape.

Application precautions

It is possible that the diversity circuit will not work on automotive radio receivers that are more than about 20 years old because of their restricted audio bandwidth. Nevertheless, the prototype was tested on several different car radios with differing ages, and it worked satisfactorily on all of them. The circuit will work with most late-model stereo receivers because they typically have as much as 20-kHz audio bandwidth. Ω

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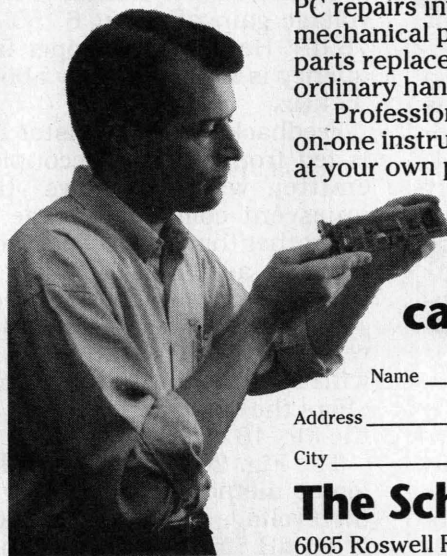
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across 5.6-kilohm emitter resistor R3. Because the emitter and collector currents of Q1 are approximately equal, the same voltage appears across 5.6-kilohm resistor R4. This sets Q1's collector at a quiescent value of two-thirds of the supply voltage. Emitter resistor R3 is AC-decoupled through C1, giving the circuit a voltage gain of 46 dB for AC signals.

Circuit variations

Figure 14 shows how the Fig. 13 circuit can be modified to give a fixed voltage gain of about 10. The operation of these circuits depends on a characteristic of common-emitter amplifiers: voltage gain equals the collector load-impedance value divided by the effective emitter-impedance value.

In Fig. 13, the effective emitter impedance equals that of the internal base-emitter junction; both equal 25 ohms at 1 milliampere. Thus this circuit has a voltage gain of about 200.

By contrast, in Fig. 14, resistor R3 is decoupled by series-connected capacitor C2 and resistor R5. Therefore, the AC emitter impedance equals the internal junction value in series with the equivalent parallel values of R3 and R5. This becomes about 560 ohms, yielding a voltage gain of about 10. Different gain values can be obtained by changing the value of R5.

Figure 15 is a simple variation of Fig. 14. In this circuit, R3 is not decoupled, and its impedance equals the value of R4. This gives the circuit a unity voltage gain. As a result, two unity-gain output signals are available: The collector output 1 is 180° out-of-phase with the input signal, but the emitter output 2 is in phase with the input signal. The circuit is called a *unity-gain phase-splitter*.

Figure 16, a variation of Fig. 11, offers 46 dB of voltage gain between the base and collector of Q1 but feedback biasing resistor R3 is AC-shunted by R2, giving the circuit a base imped-

ance of about 500 ohms.

Resistor R1 is in series with the input signal and the base of Q1. As a result, R1 in conjunction with the 500-ohm base impedance, attenuates the signal between the Q1's input and base. The overall voltage gain of the circuit is about 10, or R2/R1.

Figure 17 shows how the circuit in Fig. 11 can be modified to give wideband performance by connecting direct-coupled Q2's emitter-follower stage between Q1's collector and the output terminal. Earlier, it was pointed out that the Fig. 11 circuit can have a theoretical maximum bandwidth of 1.5 MHz. Unfortunately, the shunting effect of stray output capacitance on R2 reduces that theoretical value to about 120 kHz. However, by buffering the output through Q2, these capacitive shunting effects can be reduced, and bandwidth can be extended to several hundred kilohertz.

High-gain circuits

A single-transistor common-emitter amplifier circuit cannot have a voltage gain that is significantly greater than 46 dB when it has a resistive collector load. If voltage gain higher than 46 dB is required, the circuit must have more than one transistor.

The circuit in Fig. 18 acts like a pair of direct-coupled common-emitter amplifiers. Transistor Q1's output is fed directly into Q2's base to give an overall voltage gain of about 6,150 or 76 dB. However, the upper frequency is limited to only about 35 kHz.

Feedback biasing resistor R4 is fed from Q2's AC-decoupled emitter, which "follows" the quiescent collector voltage of Q1, rather than Q1's collector directly. In addition, the bias circuit is effectively AC-decoupled.

Fig. 19 shows an alternative version of the circuit in Fig. 18 with an PNP output stage that offers the same performance as the Fig. 18 circuit.

The Fig. 20 circuit has a different method for obtaining a high voltage gain of about 2000 or 66 dB. Transistor Q1 is connected as a common-emitter

amplifier with a split collector load of R2 and R3. Transistor Q2, connected as a common-collector amplifier or emitter follower, feeds the AC output signal from Q1's collector back to the junction of resistors R2 and R3 through capacitor C3.

This feedback *bootstraps* the value of resistor R3 (see last month's article) so that it acts as a near-infinite impedance to AC signals, and Q1 produces high voltage gain. The bandwidth of this circuit is only about 32 kHz, but its input impedance is only 330 ohms.

Other voltage amplifiers

Figure 21, for example, shows a common-base amplifier that offers wideband response. It is biased as shown in Fig. 13. However, in the Fig. 21 circuit, the base is AC-decoupled through C1, and the input signal is applied to the emitter through C3.

The circuit has a very low input impedance that equals the impedance of the forward-biased internal base-emitter junction. That impedance has the same voltage gain as the common-emitter amplifier (about 46 dB), and there is no phase shift between the input and output waveforms.

Figure 22 is a *differential* amplifier or "long-tailed" pair. The two transistors share the common-emitter resistor R7 (the "tail"), and the amplifier's bias point can be adjusted by trimmer potentiometer R8 so that identical collector currents are conducted in both transistors. That condition means that the collector voltages are equal under quiescent conditions.

Transistors Q1 and Q2 interact through R7, the emitter "tail." The output signals available from both collectors are proportional to the *difference* (the differential voltage) between the two input signals. Therefore, if identical signals are applied at the two inputs, the circuit has a zero output.

Figure 23 is the circuit of Fig. 22 modified to become a phase-splitter. It will provide two output signals 180° out-of-phase from a single-ended input. □

AN AUDIO POWER AMPLIFIER CAN boost weak signals from a tuner, CD player, or tape deck to fill a room with sound. This article focuses on the operating principles and circuitry of low-frequency power amplifiers based on the bipolar junction transistor (BJT). Recent articles in this series have discussed multi-vibrators, oscillators, audio pre-amplifiers, and tone-control circuits, all based on the BJT.

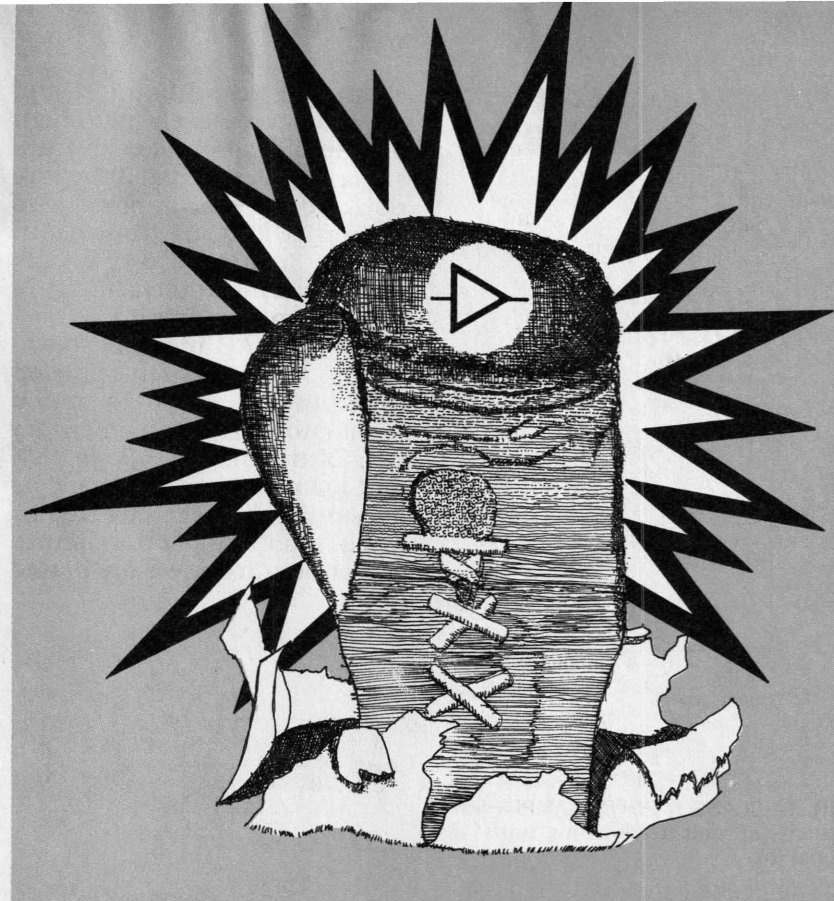
Power amplifier basics

A transistorized audio power amplifier converts the medium-level, medium-impedance AC output voltage of a preamplifier into a high-level, amplified signal that can drive a low-impedance audio transducer such as a speaker. A properly designed power amplifier will do this with minimal signal distortion.

Audio can be amplified with one or more power transistors in either of three configurations: Class A, Class B, and Class AB. Figure 1-a shows a single BJT Class A amplifier in a common-emitter configuration with a speaker as its collector load. A Class A amplifier can be identified by the way its input base is biased.

Fig. 1-a shows that BJT Q1's collector current has a *quiescent* value that is about halfway between the zero bias and cutoff positions. (The quiescent value is that value of transistor bias at which the negative- and positive-going AC input signals are zero.) This bias permits the positive and negative swings of the output collector AC current to reach their highest values without distortion. If the AC and DC impedances of the speaker load are equal, the collector voltage will assume a quiescent value that is about half the supply voltage.

The Class A circuit amplifies audio output with minimum distortion, but transistor Q1 consumes current continuously—even in the quiescent state—giving it low efficiency. Amplifier *efficiency* is defined as the ratio of AC power input to the load divided by the DC



POWER AMPLIFIERS

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power consumed by the circuit.

At maximum output power, the efficiency of a typical Class A amplifier is only 40%, about 10% less than its theoretical 50% maximum. However, its efficiency falls to about 4% at one-tenth of its maximum output power level.

A typical Class B amplifier is shown in Fig. 2-a. It has a pair of BJTs, Q1 and Q2, operating 180° out-of-phase driving a common output load, in this example another speaker. In this topology, the BJTs operated as common-emitter amplifiers drive the speaker through push-pull transformer T2. A

phase-splitting transformer, T1, provides the input drives for Q1 and Q2 180° out-of-phase.

The outstanding characteristic of any Class B amplifier is that both transistors are biased off under quiescent conditions because they are operated without base bias. As a result, the amplifier draws almost no quiescent current. This gives it an efficiency that approaches 79% under all operating conditions. In Fig. 2-b, neither Q1 nor Q2 conducts until the input drive signal exceeds the base-emitter zero-crossing voltage of the transistor. This occurs at about 600 millivolts for typical

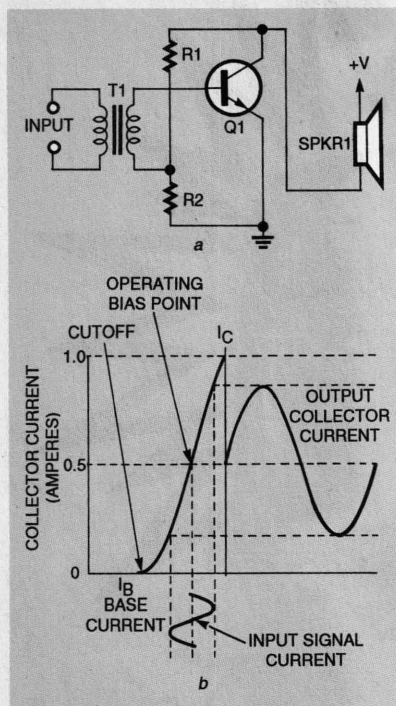


FIG. 1—CLASS A POWER AMPLIFIER circuit (a), and its dynamic transfer curve (b).

power transistor.

The major disadvantage of the Class B amplifier is that its output signal is seriously distorted. This can be seen from its dynamic transfer curve, also shown in Fig. 2-b.

Class AB fundamentals

Audio distortion caused by the crossover between two out-of-phase transistors is annoying. To overcome this defect, the classic Class B amplifier is modified into the third category called Class AB for most high-fidelity audio equipment. Fortunately, Class B distortion can usually be eliminated by applying slight forward bias to the base of each transistor, as shown in Fig 3-a. This modification sharply reduces the quiescent current of a Class B amplifier and converts it into a Class AB amplifier.

Many early transistorized power amplifiers were Class AB, as shown in Fig. 3-a, but that circuit is rarely seen today. That circuit requires one transformer for input phase-splitting and another for driving the speaker, both costly electronic components.

In addition, the electrical

characteristics of both Q1 and Q2 must be closely matched. The amplification of each transistor will be unequal if they are not, and it will be impossible to minimize output distortion. Figure 3a shows a dynamic transfer characteristic for a Class AB power amplifier.

The Class AB amplifier shown in Fig. 4 avoids both transformers and the need to match transistors. A complementary pair of transistors (Q1 an NPN and Q2 a PNP) is connected as an emitter follower. Powered by a split (dual) supply, the circuit's two emitter followers are biased

sistor Q1 acts as a current source with a very low output (emitter) impedance; it feeds a faithful unity-gain copy of the input voltage signal to the speaker. The transistor characteristics have little or no effect on this response.

Similarly, negative swings of the input signal drive Q1 off and Q2 on. Because Q2 is a PNP BJT, it becomes a current sink with minimal input (emitter) impedance. It also produces a faithful unity-gain copy of the voltage signal to the speaker, again with Q2's characteristics having little or no effect on the circuit's response.

As a result, the Fig. 4 circuit does not require that Q1 be matched to Q2, and neither input nor output transformers are required. Modification of this circuit, as shown in Figs. 5-a and b, work from single-ended power supplies. In Fig. 5-a, one side of the speaker is connected to the amplifier through high-value blocking capacitor C3, and the other end is connected to ground; in Fig. 5-b, one side is connected to C3 and the other side is connected to the positive supply. All three circuits are popular in modern high-fidelity audio

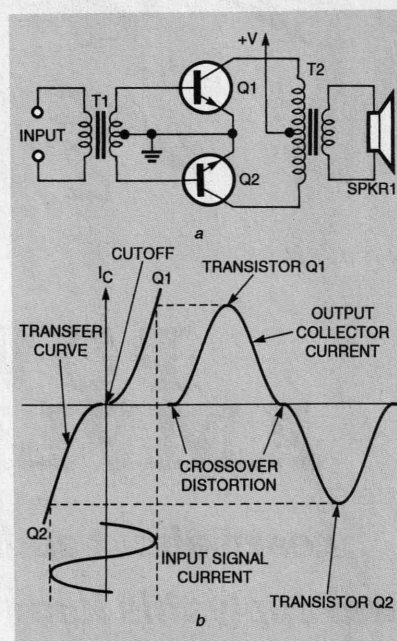


FIG. 2—CLASS B POWER AMPLIFIER circuit (a), and its dynamic transfer curve (b).

through R1 and R2 so that their outputs are at zero volts; no current flows in the speaker under quiescent conditions.

Nevertheless, a slight forward bias can be applied with trimmer potentiometer R3 so that Q1 and Q2 pass modest quiescent currents to prevent crossover distortion. Identical input signals are applied through C1 and C2 to the bases of the emitter followers, which avoid a split-phase drive.

When an input signal is applied to the Fig. 4 circuit, the positive swing drives PNP Q2 off while driving NPN Q1 on. Tran-

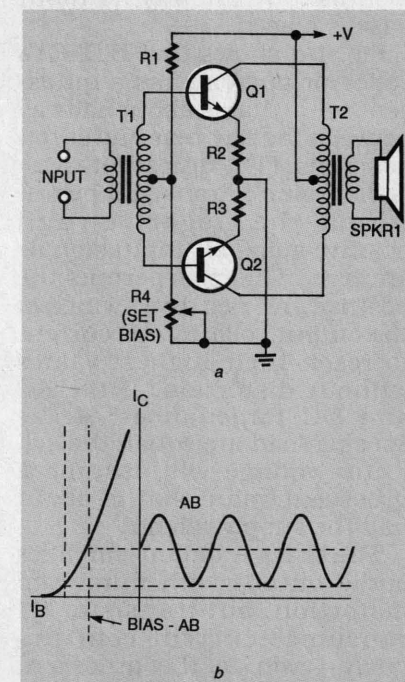


FIG. 3—CLASS AB POWER AMPLIFIER circuit (a), and its dynamic transfer curve (b).

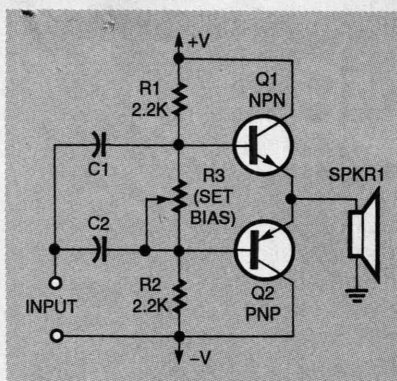


FIG. 4—CLASS AB POWER AMPLIFIER with a complementary emitter-follower output that must be powered by a dual supply.

power amplifiers based on integrated circuitry.

Class AB variations

The circuit in Fig. 4-a is a unity-voltage gain amplifier so one obvious improvement is to add a voltage-amplifying driver stage, as shown in Fig. 6. Transistor Q1, configured as a common-emitter amplifier, drives two emitter followers, Q2 and Q3, through its collector load resistor R1.

Note that Q1's base bias is derived from the circuit's output through resistors R2 and R3. This configuration provides DC feedback to stabilize the circuit's operating points and AC feedback to minimize signal distortion.

The Fig. 6 circuit illustrates how a form of auto-bias can be applied to Q2 and Q3 through the silicon diodes D1 and D2. If the simple voltage-divider biasing method in Fig. 4 is used in the Fig. 6 circuit, its quiescent current will increase as ambient temperature rises and decrease as it falls. (This is caused by the thermal characteristics of a transistor's base-emitter junctions.)

The biasing in Fig. 6 is derived from the forward voltage drop of series diodes D1 and D2 whose thermal characteristics are closely matched to those of the base-emitter junctions of Q2 and Q3. Consequently, this circuit offers excellent thermal compensation.

Practical amplifiers include a pre-set trimmer potentiometer in series with D1 and D2. This

component makes it possible to adjust bias voltage over a limited range. Low-value resistors R4 and R5 in series with the emitters of Q2 and Q3 provide some negative DC feedback.

The impedance of the Fig. 4 circuit equals the product of the speaker load impedance and the current gain of either Q1 or Q2. The circuit can be improved by replacing transistors Q1 and

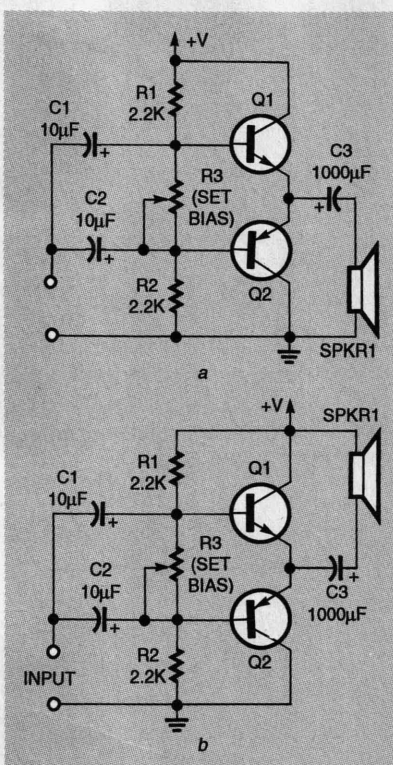


FIG. 5—ALTERNATIVE CLASS AB power amplifier that can be powered with a single-ended supply.

Q2 with Darlington pairs which will significantly increase the circuit's input impedance and increase the amplifier's collector load capacity.

Figures 7 to 9 show three different ways of modifying the Fig. 6 circuit by replacing individual transistors with Darlington pairs. For example, in Fig. 7, transistors Q2 and Q3 form a Darlington NPN pair, and Q4 and Q5 form a Darlington PNP pair. There are four base-emitter junctions between the bases of Q2 and Q4, and the output circuit is biased with a string of four silicon diodes, D1 to D4, in series to compensate for the Darlington pairs.

In Fig. 8, Q2 and Q3 are a Darlington NPN pair, but Q4 and Q5 are a complementary pair of common-emitter amplifiers. They operate with 100% negative feedback, and provide unity-voltage gain and very high input impedance. This *quasi-complementary* output stage is probably the most popular Class AB power amplifier topology today. Notice the three silicon biasing diodes, D1, D2, and D3.

Finally, in Fig. 9, both pairs Q2 and Q3 and Q4 and Q5 are complementary pairs of unity-gain, common-emitter amplifiers with 100% negative feedback. Because the pairs produce outputs that are mirror images of each other, the circuit has a complementary output stage. Notice that this circuit

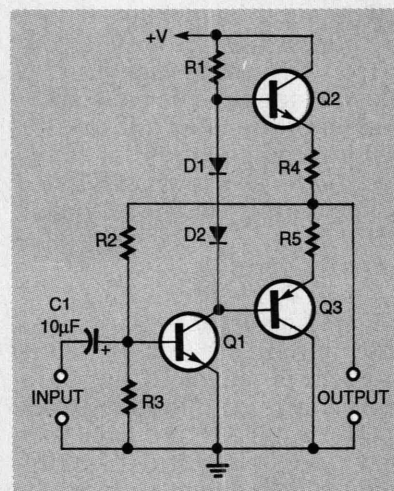


FIG. 6—COMPLEMENTARY POWER amplifier with a driver and auto-bias.

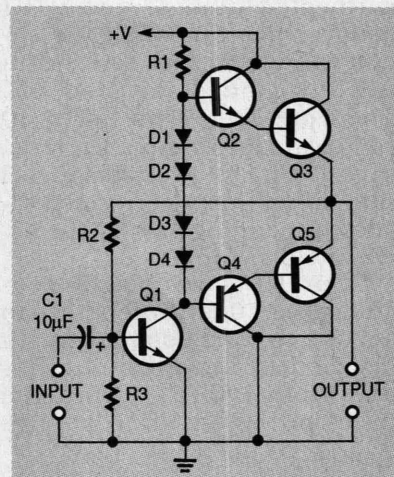


FIG. 7—POWER AMPLIFIER with Darlington output stages.

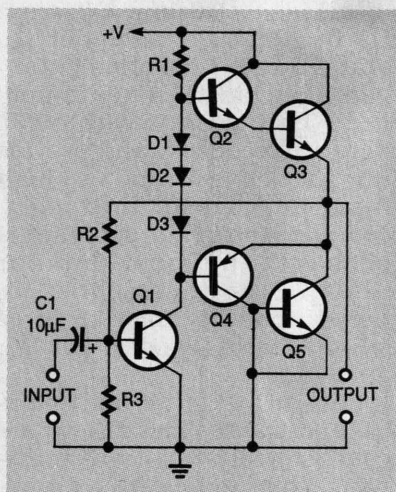


FIG. 8—POWER AMPLIFIER with partial complementary output stages.

has only two silicon biasing diodes, D1 and D2.

Amplified diodes

The circuits in Figs. 6 to 9 include strings of two to four silicon biasing diodes. Each of those strings can be replaced by a single transistor and two resistors configured as an *amplified diode*, as shown in Fig. 10.

The output voltage of the circuit, V_{OUT} , can be calculated from the formula:

$$V_{OUT} = V_{BE} \times R1 + R2/R2$$

If resistor R1 is replaced by a short circuit, the circuit's output will be equal to the base-emitter junction "diode" voltage of Q1 (V_{BE}). The circuit will then have the thermal characteristics of a discrete diode.

If resistor R1 equals R2, the circuit will act like two series-connected diodes, and if R1 equals three times R2, the circuit will act like four series-connected diodes, and so on. Therefore, the circuit in Fig. 10 can be made to simulate any desired whole or fractional number of series-connected diodes, depending on how the R1/R2 ratios are adjusted.

Figure 11 shows how the circuit in Fig. 10 can be modified to act as a fully adjustable "amplified diode," with an output variable from 1 to 5.7 times the base-emitter junction voltage (V_{BE}).

Bootstrapping

The main purpose of the Q1

driver stage in Fig. 6, the basic complementary amplifier, is to give the amplifier significant voltage gain. At any given value of Q1 collector current, this voltage gain is directly proportional to the effective Q1 collector load value. It follows that the value of resistor R1 should be as large as

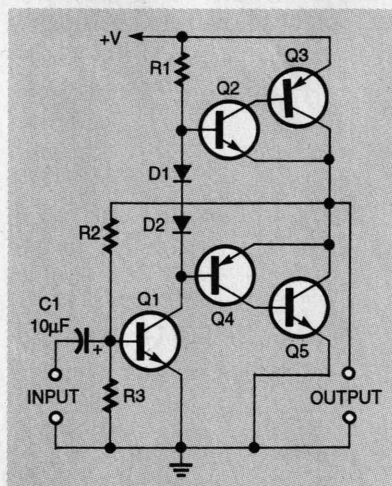


FIG. 9—POWER AMPLIFIER with complementary output stages.

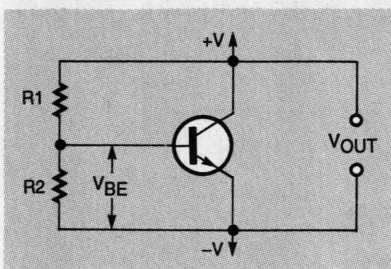


FIG. 10—FIXED-GAIN AMPLIFIED diode circuit.

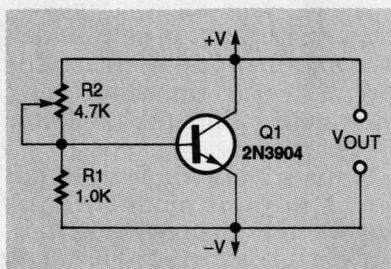


FIG. 11—ADJUSTABLE AMPLIFIED diode circuit.

possible to maximize voltage gain. However, there are several reasons why this does not work.

First, the *effective* or AC value of R1 equals the actual R1 value shunted by the input impedance of the Q2-Q3 power amplifier stage. Therefore, if R1 has a

high value, the power amplifier input impedance must be even greater. That can usually be done by replacing Q2 and Q3 with high-gain transistor pairs, as was done in Figs. 7 to 9.

The second reason is that Q1 in Fig. 6 must be biased so that its collector assumes a quiescent half-supply voltage value to provide maximum output signal swings; this condition is set by the Q1's collector current and resistor R1's value.

The true value of R1 is predetermined by biasing requirements. To achieve high voltage gain, a way must be found to make the AC impedance of R1 much greater than its DC value. This is accomplished with the *bootstrapping* technique shown in Figs. 12 and 13.

In Fig. 12, Q1's collector load consists of R1 and R2 in series. The circuit's output signal, which also appears across SPKR1, is fed back to the R1-R2 junction through C2. This output signal is a near unity-voltage-gain copy of the signal appearing on Q1's collector.

If resistor R1 has a value of 1 kilohm, the Q2-Q3 stage provides a voltage gain of 0.9. As a result, an undefined signal voltage appears at the low end of resistor R2, and 0.9 times that undefined voltage appears at the top of R2. In other words, only one tenth of the unknown signal voltage is developed across R2. Therefore, it passes

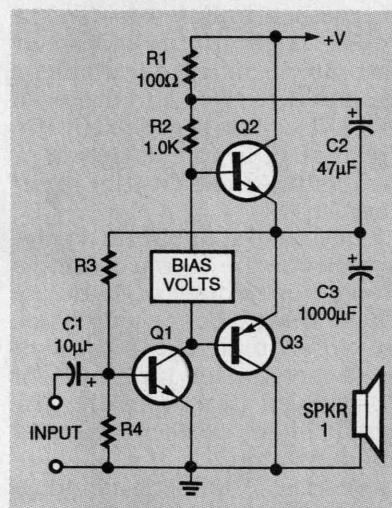


FIG. 12—POWER AMPLIFIER with a bootstrapped driver stage.

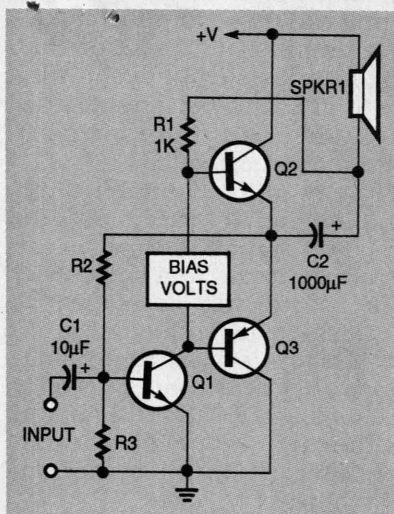


FIG. 13—ALTERNATIVE POWER amplifier with a bootstrapped driver stage.

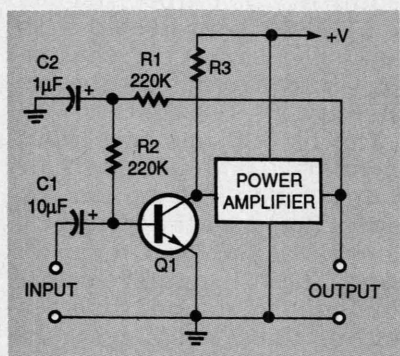


FIG. 14—DRIVER STAGE with decoupled parallel DC feedback.

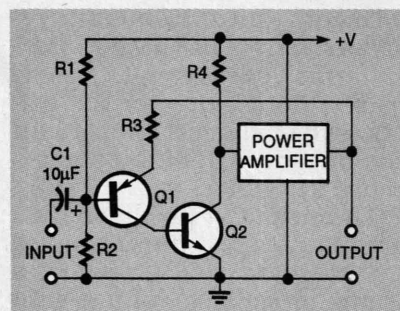


FIG. 15—DRIVER STAGE with series DC feedback.

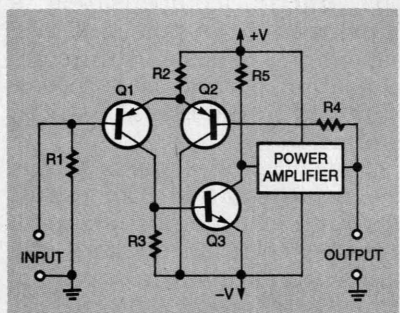


FIG. 16—DRIVER STAGE with a long-tailed pair input.

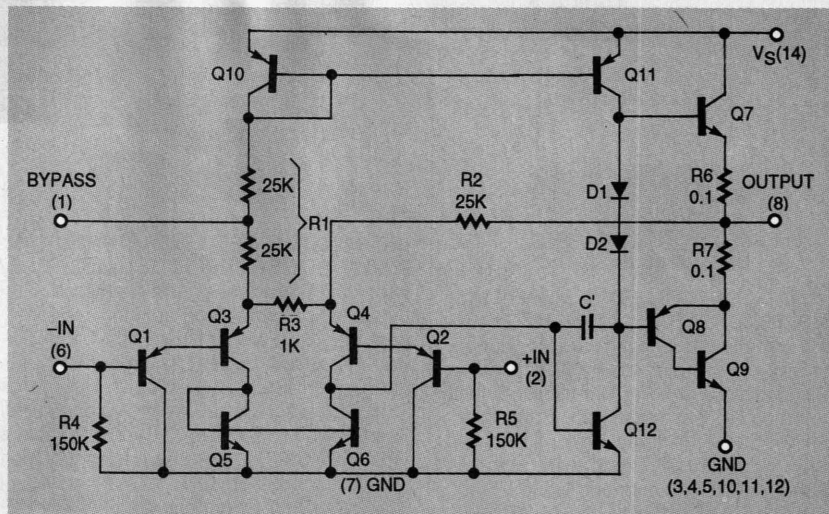


FIG. 17—SCHEMATIC FOR THE LM38 POWER AMPLIFIER IC from National Semiconductor. It has a 2-watt output rating.

only one-tenth of the signal current that would be expected from a 1-kilohm resistor.

This means that the AC signal impedance value of R2 is ten times greater (10 kilohms) than its DC value, and the signal voltage gain is increased correspondingly. In practical circuits, "bootstrapping" permits the effective voltage gain and collector load impedance of Q1 to be increased by a factor of about twenty.

Fig. 13 is the schematic for an alternative version of Fig. 12 without one resistor and one capacitor. In this circuit, SPKR1 is part of Q1's collector load, and it is bootstrapped through capacitor C2.

As an alternative to bootstrapping, the load resistor can be replaced with a simple transistor constant-current generator. This design is found in many integrated circuit audio power amplifiers.

Alternative drivers

Returning once again to Fig. 6, notice that parallel DC and AC voltage from the R1-R2 divider network is fed back to the Q1 driver stage. This is a simple and stable circuit, but its gain and input impedance are low. Moreover, it will work only over a limited power supply voltage range.

Figure 14 is a variation of the Fig. 6 circuit intended to function as a driver stage. Current feedback through resistors R1

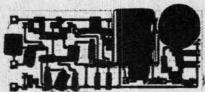
and R2 allows the circuit to work over a wide supply voltage range. The feedback resistors can be AC decoupled (as shown) through C2 to increase the gain and input impedance, but at the expense of increased signal distortion. Transistor Q1 can be replaced with a Darlington pair if very high input impedance is desired.

Another alternative driver stage, Fig. 15, depends on series DC and AC feedback to give it more gain and higher input impedance than can be obtained from the Fig. 6 circuit. In this circuit, PNP transistor Q1 is directly coupled to NPN transistor Q2.

Finally, Fig. 16 is the schematic for a driver circuit specifically intended for use in amplifiers with dual or split power supplies that have direct-coupled input and output stages referenced to ground. The input stage of this driver stage is a long-tailed pair. Both the input and output will be centered on DC ground if the values of resistors R1 and R4 are equal. This circuit is found in many integrated circuit power amplifiers.

An IC power amplifier

Improvements in the power-handling capabilities of monolithic integrated circuits have permitted power amplifiers to be integrated on a single silicon substrate or chip. The techniques for designing integrated



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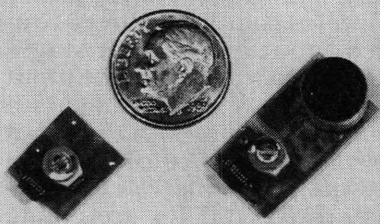


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circuit power amplifiers are similar to those for discrete device circuits. It turns out that the similarities between discrete and IC power amplifier design are closer than for most other linear circuits.

Figure 17 is a simplified circuit diagram for the LM380, an IC power amplifier, drawn in the manufacturer's data book style. The LM380 was developed by National Semiconductor Corporation for consumer applications. It features an internally fixed gain of 50 (34 dB) and an output that automatically centers itself at one-half of the supply voltage.

An unusual input stage permits inputs to be referenced to the ground or AC coupled, as required. The output stage of the LM380 is protected with both short-circuit current limiting and thermal-shutdown circuitry.

The LM380 has two input terminals. Both Q1 and Q2 are connected as PNP emitter followers that drive the Q3 and Q4 differential amplifier transistor pairs. The PNP inputs reference the input to ground, thus permitting direct coupling of the input transducer.

The output is biased to half the supply voltage by resistor ratio R1/R2 (resistor R1 is formed by two 25-kilohm resistors and R2 has a value of 25 kilohms). Negative DC feedback, through resistor R2, balances the differential stage with the output at half supply, because $R1 = 2R2$.

The output of the differential amplifier stage is direct coupled into the base of Q12, which is a common-emitter, voltage-gain amplifier with a constant current-source load provided by Q11. Internal compensation is provided by the pole-splitting capacitor C'. Pole-splitting compensation permits wide power bandwidth (100 kHz at 2 watts, 8 ohms).

The collector signal of Q12 is fed to output pin 8 of the IC through the combination of emitter-coupled Q7 and the quasi-complementary pair of emitter followers Q8 and Q9. The short-circuit current is typically 1.3 amperes. □

FIVE MORE AC CONTROL CIRCUITS will be discussed in this article. These include incandescent light dimmers and motor-speed controllers. Also discussed in this article are the basic principles of DC power control. The last four articles in this series have examined the basics of electrical and electronic power control devices and circuits, and these four articles were illustrated with schematics of semiconductor AC power switching circuits.

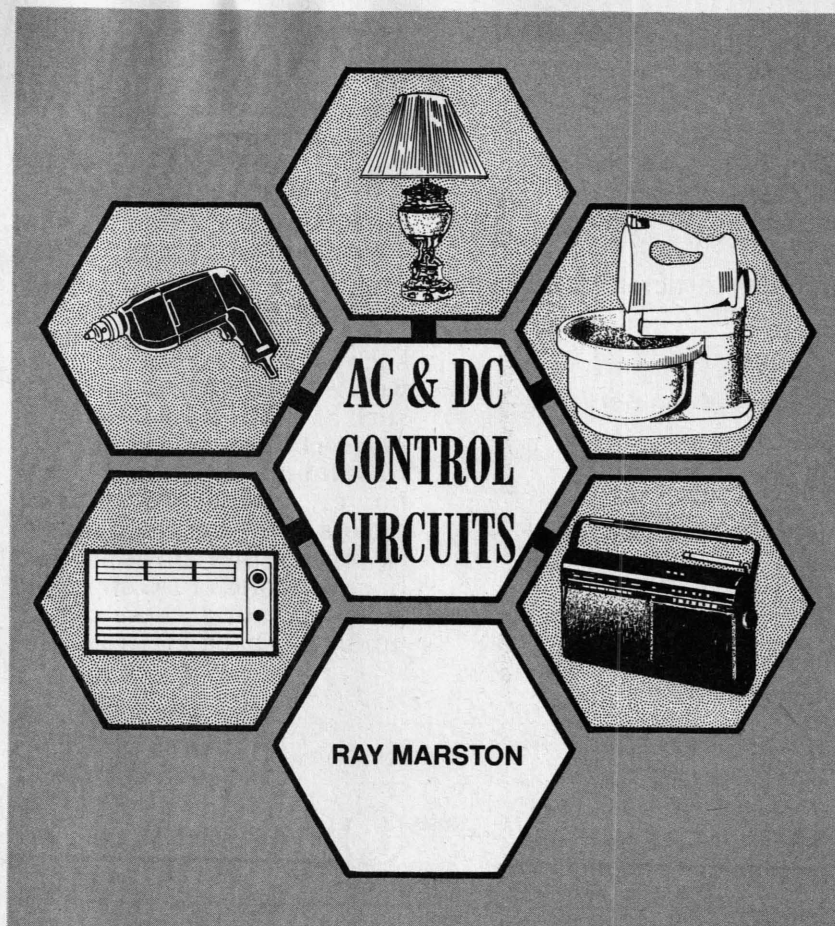
Each of the AC power control circuits presented here is based on the triac or silicon controlled rectifier (SCR) as its power switching device. All component values presented here are for switching only 120 volts AC. The reader will have an opportunity to select a triac, SCR, and diac in a rating and package style appropriate for a project or experiment.

AC light-dimming circuits

Triacs can function as efficient incandescent light dimmers because they can control the flow of current in the bulb filaments with phase-triggered power control. The triac is turned on and off once in each power-line half cycle with its duty cycle controlling the current flowing in the filament. These circuits include a simple inductive-capacitive (LC) filter power supply line to minimize radio-frequency interference (RFI).

There are three popular methods for triggering the triac by variable phase-delay methods: bilateral trigger diacs, resistor-capacitor phase-delay networks, and a line-synchronized variable-delay unijunction (UJT) trigger. Figures 1, 2 and 3 are schematics for circuits that can dim incandescent lights.

Figure 1 is the schematic for a diac-triggered incandescent light dimmer. A bilateral trigger diac is a full-wave or bidirectional thyristor. It is triggered from a blocking-to-conduction state for either polarity of applied voltage whenever the amplitude of applied voltage exceeds the breakover rating of the diac.



Learn about circuits that will control lights, AC/DC motors, and appliances as well as the fundamentals of DC power control, and put this this knowledge to work.

These devices are widely available in DO-35, axial leaded glass packages with maximum breakover voltages of 37 to 70 volts. Resistor R1, potentiometer R2, and capacitor C1 provide the variable-phase delay. An ON-OFF switch S1 is ganged to potentiometer R2 so that the lamp can be turned fully off when it is not needed.

Unfortunately, the Fig. 1 circuit exhibits control hysteresis or backlash. This can be seen if the light is dimmed almost to the point of turning it off by increasing the value of potentiometer R2 to 470 kilohms. The the lamp will not turn on again un-

til R2 is reduced to about 400 kilohms, and it then burns at a high brightness level. This backlash is caused by diac D1 partially discharging capacitor C1 each time the triac triggers. The backlash in Fig. 1 was reduced by placing 47-ohm resistor R1 in series with the diac to reduce its discharge of capacitor C1.

However, an even more effective improvement is to include a *gate-slaving circuit*, as shown in Fig. 2. The diac is triggered from C1, which "follows" the C2 phase-delay voltage but protects C2 from discharging when the diac is triggered.

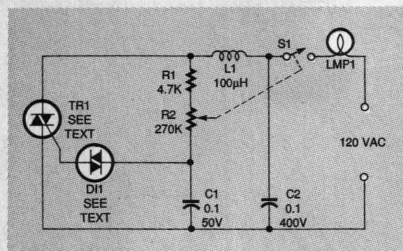


FIG. 1—SIMPLE INCANDESCENT light-dimmer circuit.

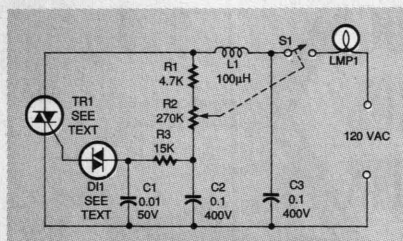


FIG. 2—IMPROVED INCANDESCENT light-dimmer circuit.

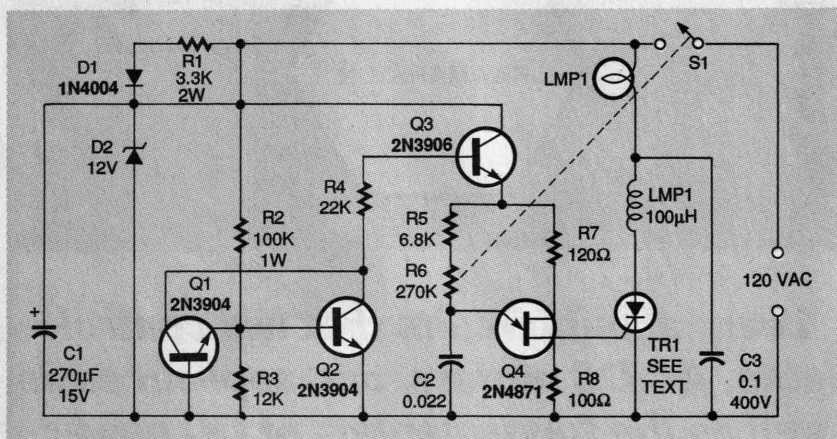


FIG. 3—ZERO-BACKLASH LIGHT dimmer with UJT triggering.

If you want a truly backlash-free circuit, install a unijunction transistor (UJT) as shown in Fig. 3. The UJT in this circuit is a 2N4871 in a TO-92 plastic package. It is powered from 12-volt DC derived from the AC line through resistor R1, diode D1, Zener diode D2, and capacitor C1.

The UJT is synchronized to the AC power line through the zero-crossing detector network consisting of Q1, Q2, and Q3, also in TO-92 plastic cases. The network turns on Q3, and it applies power to UJT Q4 at all times other than when the AC line voltage waveform approaches the zero-crossover points at the end and start of each AC half-cycle.

Thus, shortly after the start of

each half-cycle, power is applied to the UJT Q4 circuit through Q3. After a delay determined by resistor R5, potentiometer R6 and capacitor C2, a trigger pulse is applied to the triac's gate by UJT Q4. UJT Q4 resets at the end of each half-cycle, and a new sequence begins.

Universal motor control

Many consumer appliances such as food mixers and fans as well as light-duty power tools such as electric drills and sanding machines are powered by series-wound, fractional-horsepower, universal electric motors. (They are called universal because they can be powered from either AC or DC supplies). When operating, these motors produce a back electromotive

der conditions of variable load. Thus, a suitable diac with a phase-delay circuit, as shown in Fig. 4, will improve its performance. This circuit is especially useful for controlling appliances such as food mixers and sewing machines that normally operate with light loads.

By contrast, electric drills, and rotary and reciprocal sanding machines, for example, are subject to heavy load variations. Therefore, they are not suitable candidates for the circuit shown in Fig. 4. However, the alternative variable-speed regulator circuit shown in Fig. 5 is suitable. A silicon controlled rectifier (SCR) rated for 4 to 6 amperes at 200 volts is suitable as the control element that feeds half-wave power to the motor. A Motorola MCR704A1 or equivalent in a TO-220-style plastic package is suitable as SCR1.

The penalty paid for this circuit is about a 20% reduction in maximum available speed. In the off half-cycles, the back EMF of the motor is sensed by the SCR, and it provides automatic speed regulation by adjusting the next gating pulse for the SCR automatically.

The network consisting of resistor R1, potentiometer R2, and diode D2 provides only 90° of phase adjustment, so all motor pulses have minimum durations of 90°. At low speeds the circuit goes into a *skip cycling* mode, in which power pulses are delivered intermittently to suit the motor's load conditions. This circuit provides high torque under low-speed conditions.

DC power control

The remainder of this article will discuss DC power control circuits. DC power to essentially resistive loads such as incandescent lights, electric heaters, electromagnetic relay and buzz coils can be controlled by unidirectional semiconductor devices. These include bipolar power transistors, power MOSFETs, and silicon controlled rectifiers. These can be configured to give simple ON or OFF switching control or they

force (EMF) that is proportional to the motor's speed.

The effective voltage applied to universal motors is equal to the true applied voltage minus the motor's back EMF, which is directly proportional to the motor speed. Consequently, universal motors have self-regulating capability because any increase in the motor's load tends to reduce the speed and the back EMF. This, in turn, increases the effective applied voltage and causes the motor speed to increase toward its original value.

Most universal motors are made for single-speed operation. Triac phase-control circuits can provide variable speed control for these motors, but they degrade self-regulation un-

can provide variable power control, as described in earlier articles in this series.

DC power switching

The power transistor is widely used in DC power-switching applications. Figure 6 shows an NPN power transistor configured as a common-emitter amplifier. The load is positioned between the collector of Q1 and the positive power source. Transistor Q1 acts as a current sink.

This means that conventional current flows into the collector through the load. By contrast, the load in Fig. 7 is positioned between the collector of Q1 and the ground. In this schematic, transistor Q1 acts as a current source meaning that current flows from Q1's collector through the load to ground.

Common-emitter configurations offer low saturation or loss voltage (typically 200 to 400 millivolts). Their main disadvantage is that they offer low overall current and power gains (typically 100:1). These gains can be increased to 10,000:1 without increasing the saturation voltage either by cascading several common-emitter stages, as in Fig. 8, or by configuring two transistors in the *super alpha mode*, as shown in Fig. 9.

Power MOSFET can function as fast, effective DC power switches. They offer unusual characteristics and capabilities not available with bipolar power transistors. Because MOSFETs are majority-carrier devices, their switching speeds are inherently faster. Without the minority carrier-stored base charge found in transistors, storage time is eliminated. The high switching speeds allow efficient switching at frequencies above 200 kHz. This reduces the cost, size, and weight of transformers and other inductive components in switch-mode (switching) power supplies and motor controllers.

MOSFET switching speeds depend primarily on the charging and discharging of the device's capacitances, and they are essentially independent of operating temperature. The gate of a power MOSFET is electrically

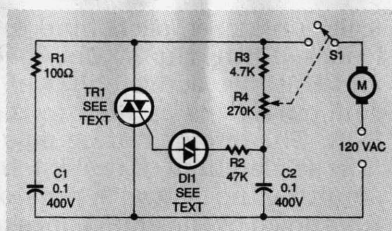


FIG. 4—SPEED CONTROL CIRCUIT for a universal AC/DC electric motor under light load.

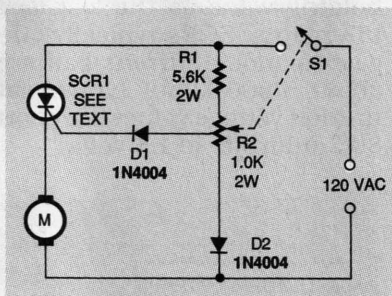


FIG. 5—SELF-REGULATING speed control circuit for a universal electric motor with a heavy load.

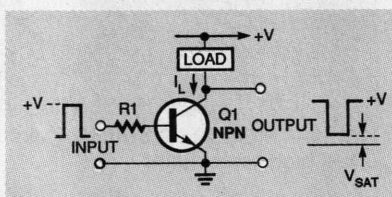


FIG. 6—THIS NPN TRANSISTOR switch circuit acts as a load-current sink.

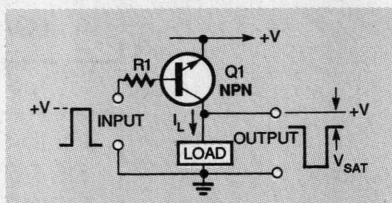


FIG. 7—THIS PNP TRANSISTOR switch circuit acts as a load-current source.

isolated from the source by an oxide layer that gives it a DC resistance greater than 40 megohms. Power MOSFET drive circuits can be relatively simple, and the gate can be driven directly from CMOS and TTL logic ICs to control high-power circuits directly.

Unlike bipolar power transistors, power MOSFETs do not require derating of power handling capability as a function of applied voltage, and destructive second breakdown does not occur if the MOSFET is operated within its specified limits.

Figure 10 is a schematic for a basic MOSFET circuit. The arrow directed toward the gate within the MOSFET schematic symbol indicates the flow of conventional current in an N-channel device and the broken vertical lines represent the source-to-drain channel, indicating that it is an *enhancement mode* or "normally off" device. A protective diode is positioned between its drain and source.

One of the more popular MOSFETs is the N-channel Motorola MTP4N50E TMOS MOSFET. The Motorola designation conveys basic specification information about the device. The first letter "M" indicates that its manufacturer is Motorola, and the "T" indicates that it is a TMOS device, Motorola's trademark for its line of power MOSFETs. The P indicates that it is in a plastic TO-220-style package. (An M, for example, would indicate a metal case, and there are letter designations for other package styles.)

The number "4" in the designation is the MOSFET's current rating in amperes. The "N" stands for channel polarity, meaning that this MOSFET is an N-channel (NPN) device. (A "P" would indicate a P-channel (PNP) device.) The 50 represents

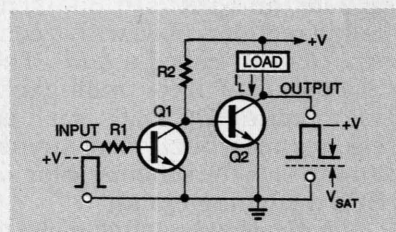


FIG. 8—HIGH-GAIN TRANSISTOR switch circuit with cascaded NPN common-emitter stages.

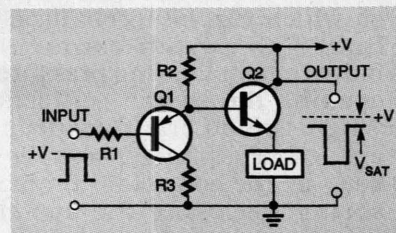


FIG. 9—HIGH-GAIN TRANSISTOR switch circuit with a super-alpha PNP pair.

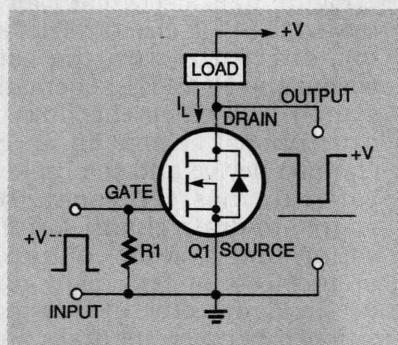


FIG. 10—POWER MOSFETs offer high-speed switching without destructive second breakdown.

the voltage rating (in volts) divided by 10, and the "E" is Motorola's symbol for its "energy-rated" devices. International Rectifier Corp. offers its equivalent trademarked HEXFET MOSFETs.

The most popular power MOSFETs today are made by the double-diffused vertical DMOS process. This geometry and process has replaced the V-groove or VMOS process widely used 20 years ago. The term "vertical" refers to the flow of current between the drain and source. For more information on these devices, refer to "Power Semiconductors" in the May 1995 *Electronics Now*.

Another effective semiconductor power switching device is the silicon controlled rectifier (SCR). As stated in the May 1995 article, an SCR is fundamentally a rectifier diode with a control element called a *gate*. The SCR is useful in controlling self-interrupting DC loads such as electric bells, buzzers, or sirens. Figure 11 is the schematic for an SCR switching circuit.

A typical load might consist of a solenoid and an activating switch in series to provide an "autoswitching" load. When the solenoid is energized, its plunger, which can be mechanically linked to switch contacts, moves against spring pressure, opening the switch contacts. When the solenoid is deenergized, the plunger is pulled back by the solenoid's internal spring, reclosing the switch contacts.

The SCR circuit of Fig. 11 provides a nonlatching load-driving because the SCR automat-

ically unlatches each time the load self-interrupts. The load and SCR are active only while gate current is applied to the SCR. The circuit can be made fully self-latching, if desired, by shunting the load with resistor R3, shown by the dotted lines to the right of diode D1.

The SCR's anode current does not fall below its minimum holding value as the load self-interrupts. SCRs typically offer gate-to-anode current gains of about 5,000:1, but typical saturation voltage values are about 800 millivolts to 1.5 volts.

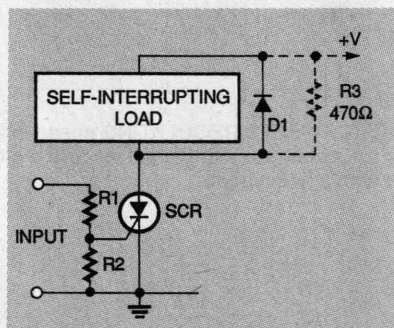


FIG. 11—THIS SCR DC switching circuit offers high power gain.

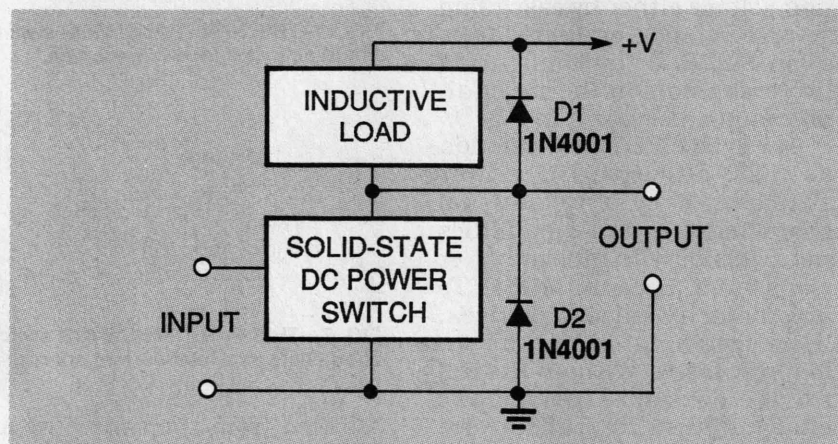


FIG. 12—SEMICONDUCTOR SWITCHING devices should be protected by damping diodes when driving inductive loads.

Handling various loads

When designing DC switching circuits, consideration must be given to the kind of load that is to be controlled and its possible damaging effects on the semiconductor switching circuitry. Here are some important points to keep in mind:

- When controlling incandescent lamps, remember that the tungsten filaments in the bulbs have a low cold resistance. The

value of the cold resistance is typically one-quarter of its value when the lamp is illuminated. Consequently, switch-on currents are typically four times greater than those that flow when the bulb is illuminated. This means that a suitable semiconductor switch for controlling 500-milliamperere lamps must have surge rating of at least 2 amperes.

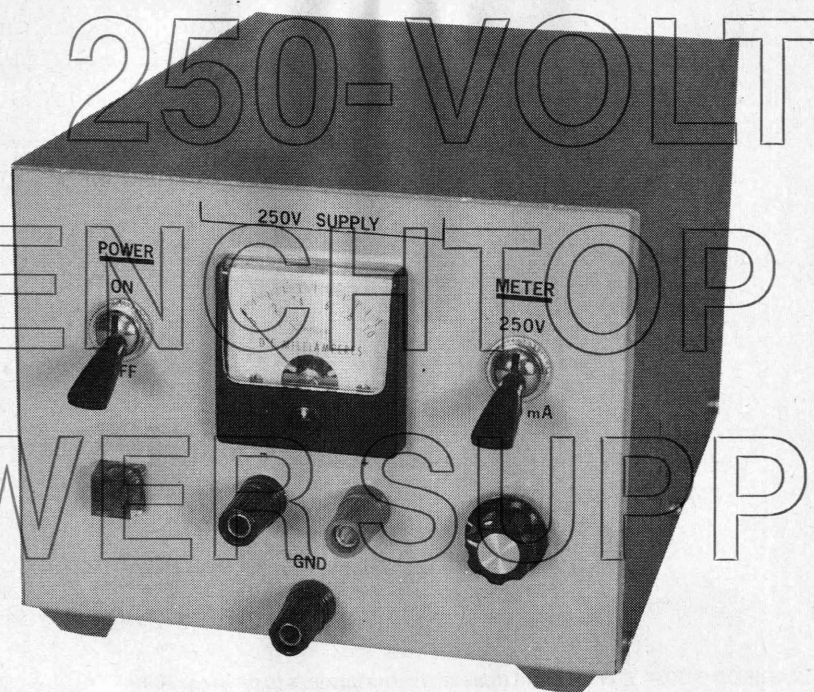
- When driving inductive loads such as relays, solenoids, speakers, and electric motors, keep in mind that these devices can generate large back EMF values when the current is switched off. Consequently, semiconductor power switches must be protected against damage from back EMF.

Figure 12 is a simplified schematic that shows how to provide circuit protection with a silicon diode that damps the back EMF. Diode D1 prevents the voltage from swinging more than a few hundred millivolts above the positive power supply value, and D2 prevents it from swinging significantly below

ground. In many applications, it might only be necessary to provide partial protection with diode D1, as shown in the SCR circuit of Fig. 11.

When driving loads that are electrically noisy (e.g., buzzers and electric motors), the loads might require damping by low value ceramic capacitors to minimize RFI generation. Also, the power supply might require ripple decoupling. □

BUILD THIS



Here's a DC power supply that makes working on high-voltage projects a breeze.

A VARIABLE HIGH-VOLTAGE POWER supply can be a great convenience to have on the bench. You can use it for servicing tube-type equipment, checking capacitors, and for general experimentation. This article describes the construction of a 0- to 250-volt, 100-milliamp regulated power supply. The supply is short-circuit protected and has an output impedance of 15 ohms. Output noise is less than 20 millivolts rms, while the temperature coefficient is only 0.03 percent per degree Celsius.

Operation

The circuit, shown in Fig. 1, uses two standard filament transformers to develop 300-volts DC. Transformer T1 steps down the 120-volt AC line power to 25 volts, and T2 steps the voltage back up to 120 volts. The 120-volt output of T2 feeds the full-wave doubler consisting of C7, C8, D1, and D2. Neon lamp NE2 and R11 form a combination bleeder and high-voltage warning light.

The heart of the circuit is IC1,

DAVID CUTHBERT

an LM317T regulator. The regulator is powered by 16 volts derived from the secondary of T1. Resistor R5 limits the current to IC1 during short-circuit operation. Current flowing through HV ADJUST potentiometer R12 generates a 1.25-volt drop across R7. If the power supply output voltage drops, IC1 turns Q1 on harder and raises the supply voltage. Resistor R3 improves the regulation by maintaining a minimum of 3 milliamps through IC1, and R2 is a parasitic oscillation suppressor for Q1.

Components R1, R4, and Q2 form the current-limit circuit, which works as follows: When 100 milliamps flows through R1, the resulting 0.7-volt drop turns Q2 on, which then steals the base drive to Q1 and limits the supply current. A 1-milliamp analog meter is used to read 250 volts or 100 milliamps full-scale. When reading voltage, the meter is connected between the emitter of Q1 and the

negative output terminal through R6, R8, and R9. When reading current, meter M1 is connected across current-sense resistor R1 through calibration trimmer R6. Diode D4 protects the supply against reverse current from the load. Switch S1-a turns power on and off, and S1-b disconnects the output when the supply is switched off.

Construction

No PC board was used to build the prototype power supply. Instead, most of the power supply circuitry is contained on a 44-pin edge-connector perforated construction board (see Fig. 2), which plugs into a matching edge connector that is mounted to the bottom of the cabinet with a couple of standoffs. The plug-in board makes the supply easier to build and to service, if that becomes necessary, in the future. However, the power-supply construction is not critical to its operation.

Power transistor Q1 must be heatsinked. The heatsink for Q1 measures 3 × 3 inches square

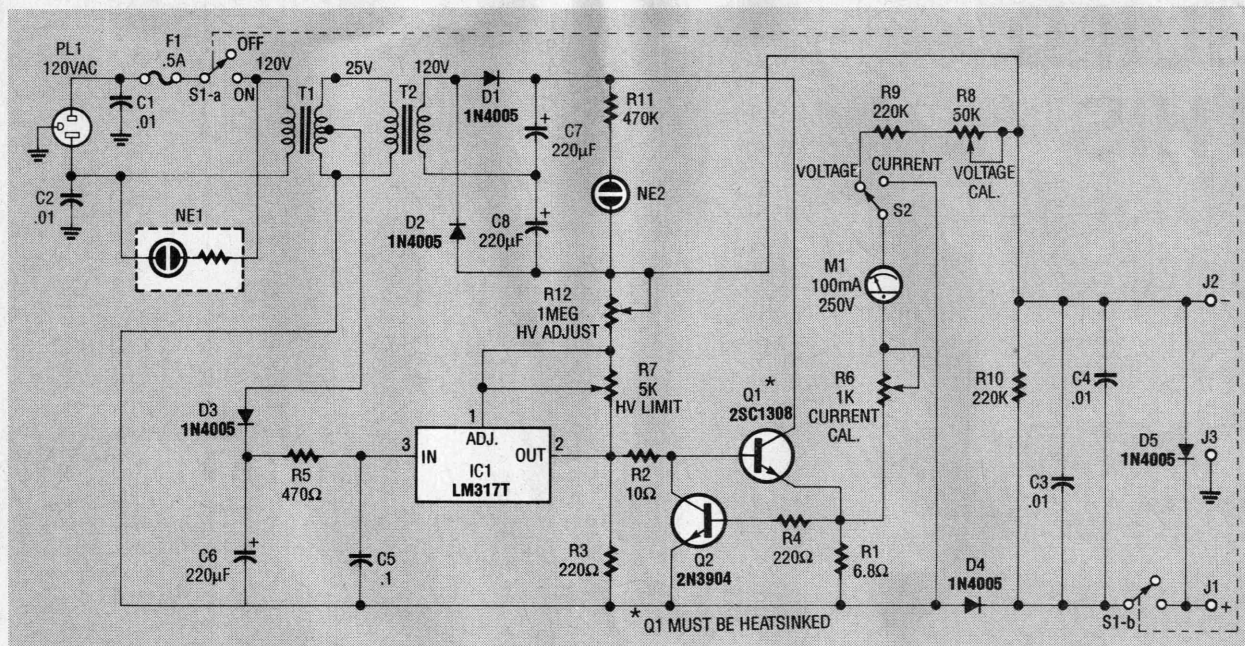


FIG. 1—HIGH-VOLTAGE DC SUPPLY. The circuit uses two transformers to develop 300-volts DC.

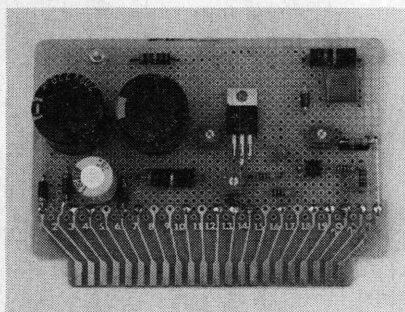


FIG. 2—MOST OF THE CIRCUITRY is contained on a 44-pin edge-constructor-perforated construction board.

with 1-inch cooling fins, and is mounted on the outside rear panel of the metal cabinet, which measures 8 × 6 × 5 inches. The case of Q1 can be more than 400 volts above ground, so make sure you mount it with sufficient insulation and put an insulating cover over Q1 to prevent a shock hazard. The fuse holder is mounted on the rear panel.

Switches S1 (POWER) and S2 (METER), which must both be rated for 250 volts AC, are mounted on the front panel of the cabinet, along with the meter (M1), HV ADJUST potentiometer R12, output jacks J1–J3, and neon indicator NE1. Diode D5 is soldered directly across jacks J1 and J2. Figure 3 shows an internal view of the prototype.

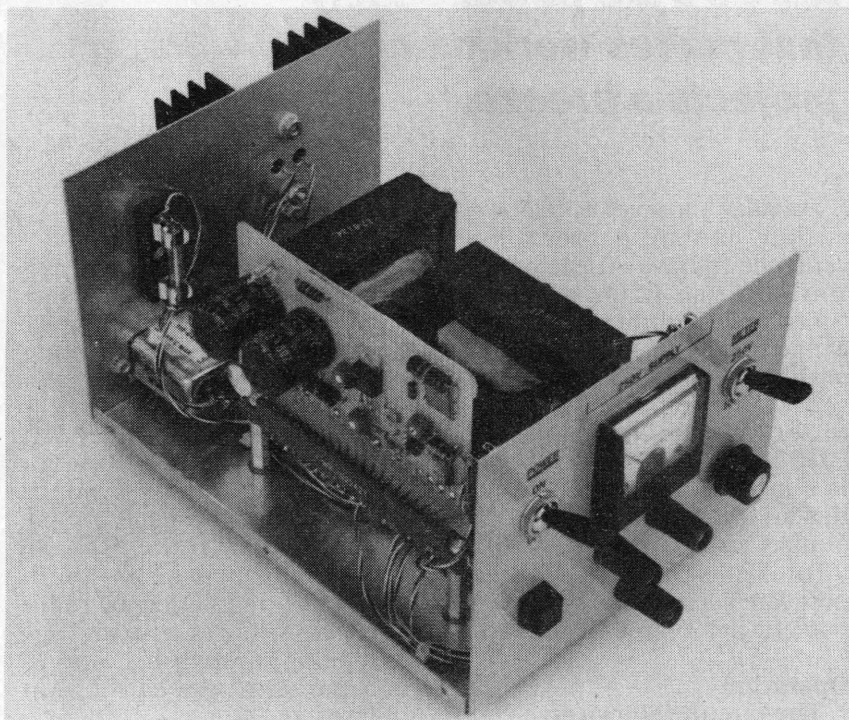


FIG. 3—PROTOTYPE POWER SUPPLY. Everything fits neatly in the metal cabinet that measures 8 × 6 × 5 inches.

Calibration

During calibration, be very careful, as 300 volts will give you quite a shock! Set your DVM to the highest DC-voltage range, and connect it to the power supply outputs. Switch the supply on, and turn the HV ADJUST knob to maximum. Adjust HV LIMIT R7 for 260 volts. Switch the supply

off, and set the DVM to the 200-milliamp range. Switch the supply on, and turn the HV ADJUST knob to maximum. The supply should limit at about 100 milliamps. Adjust the current meter calibration trimmer R6 so the current meter agrees with the DVM. Switch the DVM

continued on page 56

On-Line Tech Tips, Tools, and Test Gear for the Troubleshooter

LAST TIME WE INTRODUCED YOU TO THE BASICS OF TROUBLESHOOTING JUST ABOUT ANYTHING ELECTRONIC. THIS TIME, WE'RE GOING TO PICK UP RIGHT WHERE WE LEFT OFF AND TAKE A CLOSE LOOK AT ON-LINE TECH-TIP DATABASES, AND THEN

go on to look at the kinds of tools you really ought to have. But first, we need to deal with some old business.

It has come to our attention that due to production problems, several errors crept into the September, 1997 installment of Service Clinic. In Fig. 2, which also re-ran as Fig. 1 in October, the black and green wires on the AC plug are shown reversed; the green wire, of course, should go to the chassis ground. Also in that figure, the ground on the Magnatron should be to the chassis, and a dot is missing at the intersection between the anode of the HV diode and FA. Finally, in the middle of page 25, the value of the capacitor is given in °F; it should of course be in µF. We are sorry for any inconvenience that might have been caused by these errors and have taken steps to help prevent similar ones from cropping up in the future. Now that that's out of the way, let's move on to this month's topics.

Tech-Tips Databases

A number of organizations have compiled databases covering thousands of common problems with VCRs, TVs, computer monitors, and other electronic equipment. Most charge for their information but a few, accessible via the Internet, are either free or have a very minimal monthly or per-case fee. In other cases, a limited but still useful subset of the for-fee database is freely available.

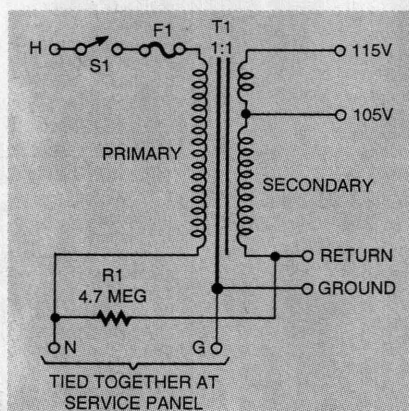


FIG. 1—HERE'S THE SCHEMATIC FOR A TYPICAL ISOLATION TRANSFORMER. It protects you against shock hazards when working on "hot chassis" equipment.

A tech-tips database is a collection of problems and solutions accumulated by the organization providing the information or other sources based on actual repair experiences and case histories. Since identical failures often occur at some point in a large percentage of a given model or product line, checking out a tech-tips database might quickly identify your problem and solution.

By using a tech-tips database, you can often simplify your troubleshooting or at least confirm a diagnosis before ordering parts. My only reservation with respect to tech-tips databases in general—and this has nothing to do with any one in particular—is that symptoms can sometimes be deceiving. A solution that works

in one instance may not apply to your specific problem. Therefore, an understanding of the hows and whys of the equipment along with some good old-fashioned testing is highly desirable to minimize the risk of replacing parts that turn out not to be bad. In simpler words, use these databases as an assistant, not as a replacement for logical troubleshooting techniques.

Another disadvantage to the databases is that you do not learn much by just following a procedure developed by others. There is no explanation of how the original diagnosis was determined, or what might have caused the failure in the first place. Nor is there likely to be any list of other components that might have been over-stressed by the original problem, and that might fail in the future because of it. Knowing that you need to replace "Q701" and "C725" to get this specific piece of gear going again in this specific instance is fine for now, but that "knowledge" won't help you to repair a different model or a different problem in the future.

One alternative to tech-tips databases is to search at <http://www.dejanews.com/> or the sci.electronics.repair Usenet news group for postings with keywords matching your model and problem. Having said that, here are three tech-tip sites for computer monitors, TVs, and VCRs:

- <http://www.anatekcorp.com/techforum.htm> (Free)
- <http://www.electronix.com/elexcorp/techsweb.html> (\$8/month)
- <http://elmswood.guernsey.net/> (Free, currently very limited)

The following source is just for monitors. Some portions are free but others require a \$5 charge; however, that charge could get you a personal reply from a technician experienced with your

monitor, so it could be well worth it:

- <http://www.netis.com/members/bcollins/monitor.htm>

Some free monitor repair tips are also available at:

- <http://www.kmrtech.com/>

Tech-tips of the month and "ask a wizard" options are available at:

- <http://members.tripod.com/~ADC/C/> — (Home page)

- <http://members.tripod.com/~ADC/C/tips.htm> — (Tech-tips of the month)

The following is specifically for microwave ovens. In addition to a large database of specific repairs, there is a great deal of useful information and links to other sites:

- <http://www.yup.com/microtech/>

Hand Tools

Invest in good tools. If you are into garage sales, you can often pick up excellent, well-maintained tools very inexpensively, but be selective—there's a lot of junk out there. In the end, substandard tools will slow you down and prove extremely frustrating to use. Keep your tools healthy—learn to use a whetstone or grinding wheel where appropriate (screwdrivers, drill bits, etc.) and put a light film of oil (e.g., WD40) on steel tools to prevent rust.

Some of the basic hand tools you will need to accumulate include:

- Standard screwdrivers of all types and sizes including straight, Philips, Torx. You'll also need a notched straight blade

for VCR mechanical tracking adjustment—you can make one out of a standard screwdriver or buy one.

- Jewelers' screwdrivers—both straight and Philips. These are generally inexpensive but quality usually varies in direct proportion to the price.

- Small socket driver set.

- Security bits for some video games, PS2s, etc.

- Hex key wrenches or hex drivers. You'll need miniature metric sizes for VCRs.

- Pliers—long nose, round nose, curved. Both smooth and serrated types are useful.

- Adjustable wrench (small).

- Cutters—diagonal and flush.

- Linesman's pliers.

- Wire strippers—fixed and adjustable.

- Crimp tool.

- Alignment tools—at least a standard RCA type for coils.

- Files—small set of assorted types including flat, round, square, and triangular.

- Dental picks—useful for poking and prodding in restricted areas (but you knew that).

- Locking clamps and or hemo-stats—for securing small parts while soldering, etc.

- Magnetic pickup tool—you can never tell when you will drop something deep inside a VCR. If you keep a strong magnet stuck to your workbench, you can use it to magnetize most steel tools such as screwdrivers. Just keep anything magnetized away from the tape path and magnetic heads (and magnetic media!).

- Hand drill, electric drill, drill press—one or all. A small bench-top drill press (about 8 inches) is invaluable for many tasks. A good set of high-speed bits (avoid the 1000 bits for \$9.95 variety). Also, miniature bits for PCB and small plastic repairs are likely to be needed.

- Soldering and de-soldering equipment. You don't need a fancy rework station; a 25-watt iron and hand de-soldering pump will be adequate for most tasks.

Basic Test Equipment

Obviously, you can load up on exotic test equipment, but it is far from required. What's listed here are those instruments that are most used. You might at first not consider all of what follows to fit the category of test equipment, but an old TV, for example, can often provide as much or more useful information about a video signal than a fancy waveform analyzer:

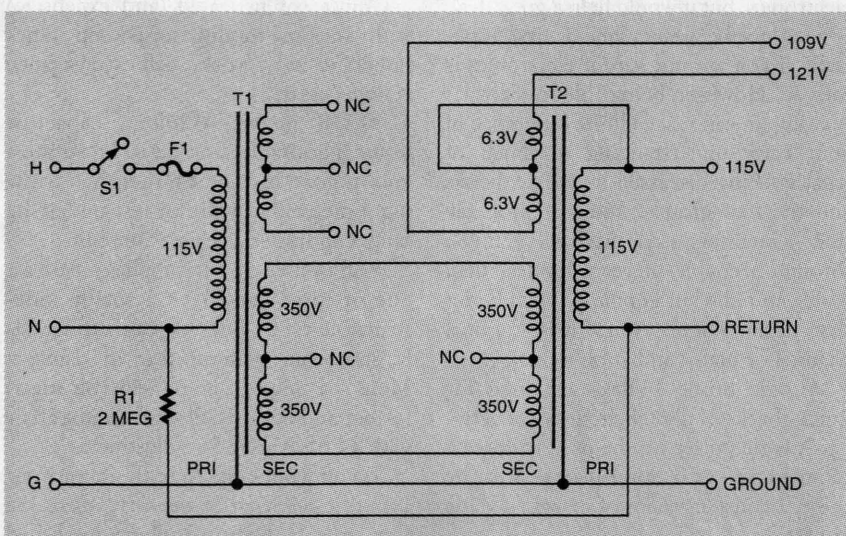


FIG. 2—YOU CAN BUILD YOUR OWN ISOLATION TRANSFORMER using back-to-back power transformers salvaged from old tube-type TV receivers.

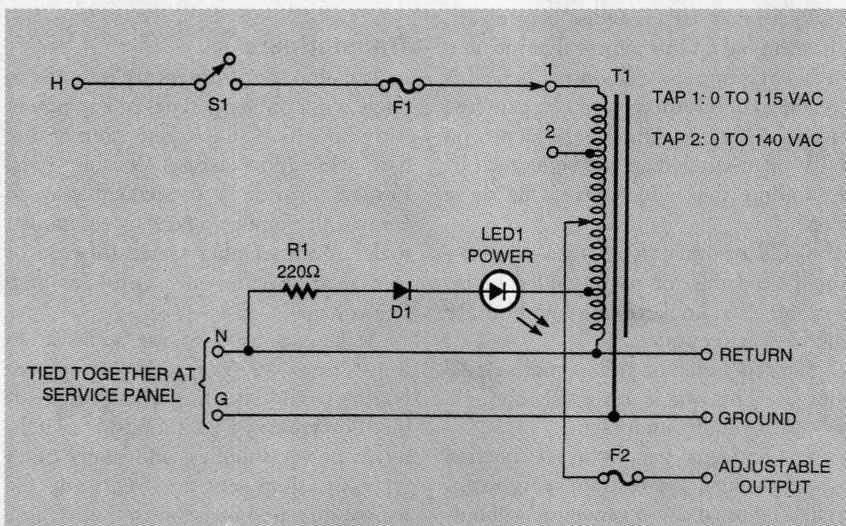


FIG. 3—THE INTERNAL WIRING OF A TYPICAL VARIAC is shown here. A VARIAC lets you vary the AC line voltage that is input to a piece of equipment, something that is a very helpful asset when troubleshooting.

•DMM and/or VOM—I prefer to have both. A good old Simpson 260 analog meter is better in many ways than a cheap digital multimeter. For most measurements, I still use a 25-year-old Lafayette (remember them?) VOM. I only go for the DMM when I need to measure really low resistances or where better accuracy is needed (though that can be deceptive). Just because a DMM has $3\frac{1}{2}$ digits does not mean it is that accurate. Check the manual, it may prove enlightening. The Simpson 260 also has a nice 5000-volt AC/DC scale that many newer digital instruments lack.

Scales for transistors, capacitors, a frequency counter, etc. are not really essential. A diode-test function on a DMM is needed, however, to properly bias semiconductor junctions. However, even that is not useful for in-circuit tests or for some power transistors or transistors with built-in damper diodes and/or base resistors.

Make sure you have a good well-insulated set of test probes. This is for your own safety as you may be measuring relatively high voltages. Periodically inspect those for damage and repair or replace as needed. If the probes that came with your multimeter are substandard—flimsy connectors or very thin insulation—replace them as well.

A high-impedance high-voltage probe is sometimes useful for TVs and monitors. You can build one that will suffice for most consumer-electronics work.

•AC clamp-on ammeter—This tool permits the measurement of currents in appliances or electrical wiring without having to cut any wires. At most, you will need an easily constructed adapter to permit access to a single conductor of a line cord. Some multimeters offer this as an option.

•Oscilloscope—You'll need a dual trace, 10- to 20-MHz minimum vertical bandwidth unit; also one with delayed sweep is desirable, but not essential. Make sure you obtain a good set of proper $10\times/1\times$ probes. High vertical bandwidth is desirable but most consumer electronics work can be done with a 10-MHz scope. If you get into digital debugging, that is another story—bandwidths of 100 MHz and up will be required. If money is no object, buy a good digital storage scope. You can even get relatively inexpensive scope cards for PCs, but unless you are into PC-controlled instrumentation, a stand-alone scope is much more useful.

If money is a concern (or perhaps even if it isn't) consider one of the old "war horses" you often see on the surplus/used market. You will usually get more scope for your money and these things last almost forever. My "good" scope is the militarized version (AN/USM-281A) of the Hewlett-Packard 180 lab scope. It has a dual-channel 50-MHz vertical plug-in and a delayed-sweep horizontal plug-in. I have seen these going for under \$300 from surplus outfits. Other types of plug-ins are available as well. For a little more money (\$400-\$700 on the surplus/used market), you can get a Tektronix 465 100-MHz scope that will handle all but the most demanding tasks.

You don't absolutely need an oscilloscope when you are just starting out in electronics, but it would help a great deal. It need not be a fancy one at first, especially if you are not sure if electronics is for you. However, being able to see what is going on can make all the difference in your early understanding of much of what is being discussed in the textbooks and the newsgroups. You can probably find something used that will get you through a couple of years for less than \$100. An oldie but goodie is much better than nothing at all even if it isn't a dual channel or high-bandwidth model.

•Logic probe—These are used for quick checks of digital circuitry for activity. A logic pulser can be used to force a momentary 1 or 0. Some people swear by these. I consider them of marginal value at best.

•TV set and/or video monitor—One of these, and preferably a color unit, is needed for testing video equipment like VCRs, camcorders, laserdisc players, etc. I have an old CGA monitor that includes an NTSC input as well on my bench. A great deal of information can be gathered more quickly by examining the picture on a TV or monitor than can be learned by examining the video waveform on a scope.

•VCR or other video-signal source—You'll need this for testing video monitors and TVs. Look for one with both RF and baseband outputs.

•Stereo tuner or other audio-signal source—This one is, obviously enough, for testing audio equipment.

•Audio signal generator—A function generator (sine, square, triangle) is nice as well. The usual audio generator will output from a few Hz to about 1 MHz.

•Audio amplifier—The input should be selectable between line level and mic

level, and be brought out through a shielded-cable to a test probe and ground clip. This is useful for tracing an audio circuit to determine where a signal is getting lost. The amplifier's output should be connected to a loudspeaker.

•Signal injector—A readily accessible portable source of a test tone or other signal (depending on application) that can be introduced into the intermediate or early stages of a multistage electronic system. For audio, a simple transistor or 555 timer based battery-powered oscillator can be built into a hand-held probe. Similar devices can be built for RF or video testing.

•RF signal generator—This is needed for serious debugging of radio and tuner front-ends. These generators can get quite sophisticated (and expensive) with various modulation/sweep functions. For most work, such extravagance is unnecessary.

•LCR meter—While a capacitor tester is desirable, I prefer to substitute a known good capacitor rather than trusting a meter that will not test under the same conditions as exist in-circuit.

•Adjustable power supplies—At least one of these should be a totally indestructible type—one you can accidentally short out without fear of damage. Mine is a simple 1 amp 0-40-volt transformer and rectifier/filter capacitor affair with a little Variac for adjustment.

If you would like to try building your own test gear, *Test Equipment Projects You Can Build*, by Delton T. Horn, published by Tab Books, a division of McGraw-Hill, Inc., 1992 has a number of simple projects you can try.

Transformers

Isolation transformers are essential to safely work on many types of equipment with exposed AC line connections or that have a live (hot) chassis. Variable transformers provide a convenient way to control the input voltage to equipment to determine whether a fault still exists or to evaluate performance at low or high line voltage.

Make it a habit to use an isolation transformer to power the equipment you are troubleshooting whenever possible. Portions of TVs, monitors, switch-mode power supplies, and many other types of equipment are frequently fed from a direct connection to the AC line without a power transformer (which would provide the isolation function). The DC power rails will typically be

between 150 and 300 volts and can deliver momentary current of potentially lethal multiple amps!

Since earth ground and the power line neutral are connected together at your service panel (fuse or circuit-breaker box), grounds like cold water pipes, test equipment chassis, and even a damp concrete floor make suitable returns for the line voltage (hot or live wire). Since this is equally true whether the conductor is a wire or your body, such a situation is very dangerous. An isolation transformer as its name implies provides a barrier that protects against accidental contact with an earth ground. With the transformer in place, such contact results in negligible current flow (mainly due to the parasitic capacitance of the transformer)—a slight tingle at worst. That also protects your test equipment as well as the device you are troubleshooting, since without an isolation transformer, a similar accidental contact could result in a short circuit, sparks, destroyed parts, etc.

Figure 1 shows the schematic of a typical isolation transformer. Note that the ground is included on the secondary side. That is actually needed for safety with certain types of equipment like microwave ovens where the HV return is to the chassis. Most other consumer gear will only have a 2-wire cord and do not use the ground.

Even though the power line neutral and ground wires are tied together at the main service panel (fuse or circuit-breaker box), the transformer prevents any significant current flow between any of its outputs and earth ground should a fault occur. The resistor in Fig. 1 permits any static charge to leak off to ground. Since it is quite large—several megohms—no perceptible current flows between the secondary and primary sides, but that value is still low enough to dissipate any static charge. **CAUTION:** The resistor must be a high-voltage rated type (as in 4200 volts, isolation, large-size, light-blue color) to assure that arc over will not result due to voltage differences that may be present when the isolation transformer is being used in its normal manner.

Isolation transformers can be purchased or you can make your own out of a pair of similar power transformers connected back to back. I built mine from a couple of old tube-type TV power transformers mounted on a board with an outlet box. Their high-voltage secondary windings were connected together. The

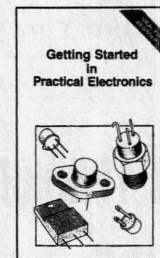
unused low-voltage secondary windings can be put in series with the primary or output windings to adjust voltage. The schematic is shown in Fig. 2. Note that there should be a fuse in the primary to protect against faults in the transformer as well as the load. Use a slow-blow type. The inrush current of the transformer will depend on the part of the cycle when the switch is closed (worst is actually near the zero crossing) as well as the secondary load. Though not shown in the schematic, it is a good idea to also place a fast-blow fuse in series with the secondary to protect the load. However, the inrush current of the degaussing coils in TV sets and monitors will often pop a normal or fast-blow fuse when no actual problems exist. (It is probably a good idea to disconnect the degaussing coils while testing unless they are suspected of being the source of the problem.)

A variable autotransformer (also known as a Variac, which is the trade name of one manufacturer) doesn't need to be large—a 2-amp unit mounted with a switch, outlet and fuse will suffice for most tasks. However, a 5-amp or larger Variac is desirable. If you will be troubleshooting 220-VAC equipment in the US, there are Variacs that will output 0 to 240 VAC from a 115-VAC line. As valuable as a Variac can be, it is important to remember that a **Variac is NOT an isolation transformer!** Don't make the mistake of using one in place of an isolation transformer. Note that there are combination units, also known as variable isolation transformers. If you have one, great; but if not, there is no need to buy such a combination unit. A Variac followed by a normal isolation transformer will work fine.

Figure 3 shows the internal wiring of a typical Variac. Note that there is no isolation as the power-line neutral and ground are tied together at the main service panel (fuse or circuit-breaker box)! Also note that the "Power-LED circuit" in Fig. 3 is soldered directly to a winding location that has been determined to produce about 6 VAC.

Next time I'll continue this discussion by presenting my thoughts on some basic ancillary equipment every troubleshooter ought to have, as well as some hints on where and how to get service literature including schematics. In the meantime, why not visit my sci.electronics.repair FAQ site on the Internet at www.repairfaq.org. You can reach me directly via e-mail at sam@stdavids.picker.com. **EN**

You can Build Gadgets! Here are 3 reasons why!

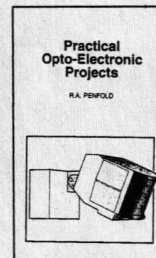


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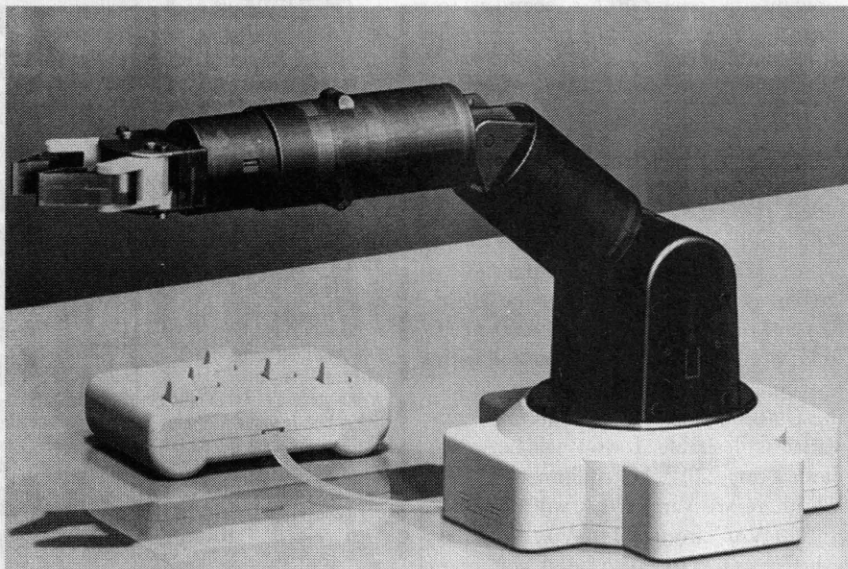
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The Robotic Arm Trainer Kit from OWI teaches the basic principles of robotic sensing and locomotion. It includes a five-switch wired controller that permits the arm to grab and release, lift, lower, rotate both the wrist and the base, and pivot sideways 120 degrees. Five motors and five joints allow for flexibility and fun.

After the robotic kit is assembled, the dynamics of the gear assembly can be observed through the transparent arm. The state-of-the-art gear mechanism is a combination of link-mechanism with a motor gear-box control.

The robotic arm has a maximum length of 18 inches outwards and a maximum height upwards of 14 inches. It uses 4 "D" batteries (not included) as a power source and has a suggested selling price of \$69.95

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Mini-Stick Meters

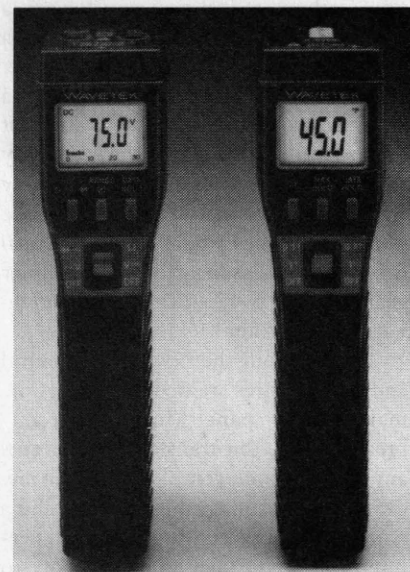
Wavetek has introduced two uniquely shaped digital multimeters, the ST75 and TM45. Traditional handheld DMMs must be put down by the technician while holding the test leads. The "mini-stick" design of these meters, however, allows the user to simultaneously hold the meter safely and read the display.

Designed especially for electricians, plant and maintenance engineers, and others who need a compact tool, the small thin style of the meters makes it easy to carry in tool belts and helps it fit easily in tool boxes (it measures approximately 6³/₄-inches high by 1⁵/₈-inches wide at its widest). Wavetek's ST75 and TM45 are very useful for applications where space is tight, for quick tracing on wiring panels and circuits, and for blower and motor circuit troubleshooting.

The ST75 is a complete volt/ohm stick DMM, featuring digital and analog bargraph display, 3200-count resolution, autoranging, data hold, quick continuity

checking, and diode testing. It measures DC and AC voltage up to 600 volts and resistance up to 32 megohms.

The TM45 is a digital thermometer, geared to commercial and industrial applications, with a wide temperature range up to 2000° Fahrenheit and 1300° Celsius. Features include temperature measurement that is switchable between Fahrenheit and Celsius, compatibility with Type K-thermocouples, and data and maximum display hold.



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RAY MARSTON

SINEWAVES CAN BE PRODUCED BY synthesizing them from either digital or analog waveforms. This second article on sinewave generation is about sinewave synthesizers, function generators, and inductance capacitance (LC) oscillator circuits. It picks up from last month's article about resistance-capacitance (RC) sinewave generators that focused on Wien-bridge and twin-T oscillators.

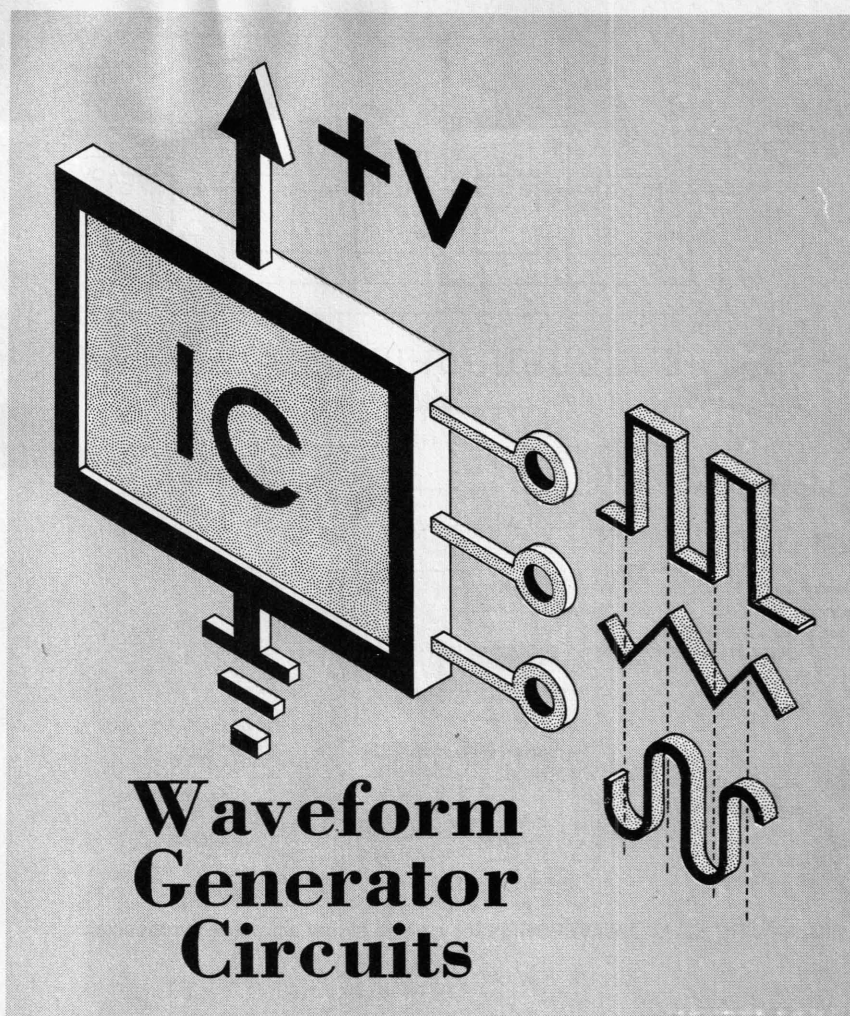
Sinewave synthesizers

Figure 1 is a block diagram of a 1-kHz digital sinewave synthesizer comprised of five D-type flip-flop circuits in a Johnson counter. It is accompanied by the waveforms that would be obtained at four different locations, A, B, C, and D. The figure illustrates how a sinewave can be generated digitally by refining a coarse sinewave in a series of digital steps and then filtering out all of the digital signal's high-frequency components.

In this example, a clock signal is fed to the input of a five-stage Johnson counter. Four of the counter outputs are added together with a resistor network to produce a coarse sinewave which is then converted into a reasonably pure waveform by low-pass filter capacitor C1.

The sinewave output frequency of this circuit is one-tenth of its input clock frequency. To understand this circuit, recall that digital signals generate only odd harmonics. Thus, the harmonics that must be eliminated to obtain a reasonable sinewave are the ninth, eleventh, nineteenth, twenty-first and higher. These are easily filtered out by capacitor C1.

Figure 2 is a simple, low-cost, 1-kHz digital sinewave synthesizer based on a CD4018B CMOS presettable divide-by-N counter. This CMOS monolithic integrated circuit, designated IC1, was developed by RCA but is now produced by Harris Semiconductor and other alternate sources. The sinewave synthesizer also contains Q1, a 2N3904 NPN transistor, that



Learn about waveform generation with synthesizers, integrated-circuit function generators, and LC oscillators, and apply that knowledge to your projects.

converts an external 10-kHz input signal which is compatible with the signal on CLOCK pin 14.

The lowest significant harmonic of this circuit's 1-kHz output is the ninth harmonic, at -36 dB relative to the fundamental frequency. (This sinewave has about 2 % distortion.) If capacitor C1 is replaced by a second-order low-pass filter, all harmonics are attenuated 65 dB with respect to the fundamental, resulting in high-quality sinewaves with about 0.1 % distortion.

Sinewaves can also be synthesized from linear triangular waveforms. Integrated circuits,

available from many sources, will perform this function. One example is the ICL8038, a precision waveform generator made by Harris Semiconductor and alternate-sourced by four other manufacturers. Another example is the XR-2206, made by Exar Corporation and alternate-sourced by at least three other manufacturers.

The Harris ICL8038

The ICL8038 waveform generator can produce accurate sine, square, triangular, sawtooth, and pulse waveforms with a minimum of external components. The frequency or repeti-

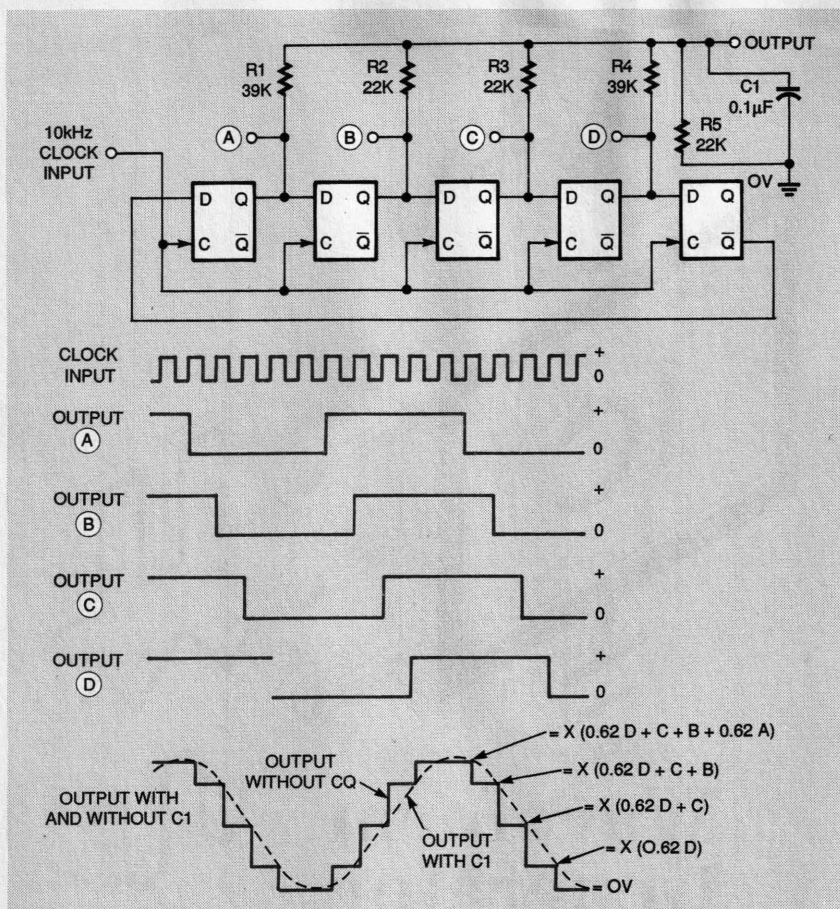


FIG. 1—CIRCUIT AND WAVEFORMS for a 1-kHz digital sinewave synthesizer.

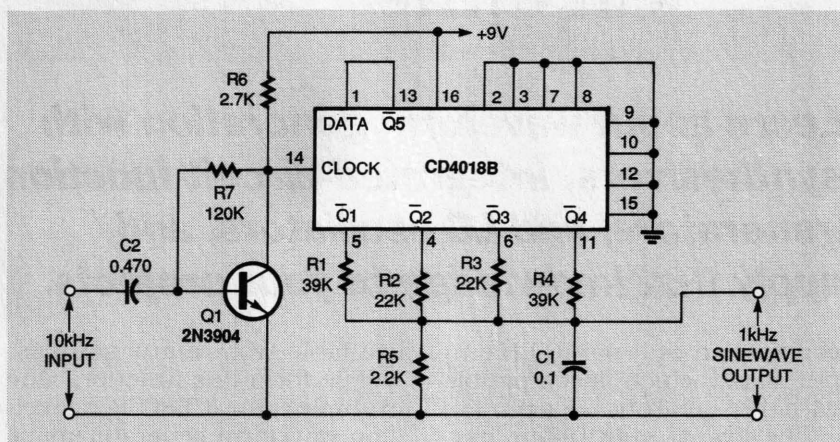


FIG. 2—A 1-kHz DIGITAL SINEWAVE synthesizer circuit.

tion rate can be selected externally from 0.001 Hz to more than 300 kHz with either resistors or capacitors. An external voltage will permit frequency modulation and circuit sweeping.

Figure 3 is the functional block diagram for the ICL8038. It contains two current sources (No. 1 and No. 2), two comparator circuits (No. 1 and No.

2), and a flip-flop. Square waves are obtained from a buffer fed by the flip-flop, and a second buffer is connected on the input sides of the parallel comparators to provide triangle waves. Sinewaves are obtained by passing the triangular waves through a sine converter.

Figure 4 is the pin configuration diagram for the ICL8038. It can be powered either from a

single power supply (10 to 30 volts) or a dual power supply (± 5 to ± 15 volts). The device provides sinewaves with only 1 % distortion and triangular waves with 0.1 % linearity.

Figure 5 is a schematic showing the ICL8038 configured as a sine/square/rectangle-wave generator powered from a single supply. Notice that in this diagram, the FM BIAS pin 7 is short-circuited directly to the FM SWEEP INPUT pin 8. With this configuration, the values of capacitor C1 and resistors R_A and R_B determine operating frequency. The resistors set the operating values of the two internal current sources that alternately charge and discharge external capacitor C1.

Refer again to the functional diagram Fig. 3. Current source No. 2 is switched on and off by a flip-flop, while current source No. 1 remains on continuously. Without going into detail on circuit function, the important point here is that external capacitor C1 charges at a rate set by R_A until its voltage reaches two-thirds of the supply voltage. At that time, the flip-flop is triggered and changes state. External capacitor C1 is discharged linearly at a rate set by R_B until its voltage falls to one-third of the supply voltage. The flip-flop is then triggered to change state again and the whole process repeats again.

Resistors R_A and R_B can have values between 1 kilohm and 1 megohm. If these values are equal, the generator circuit runs at a frequency of $0.3/RC$. It generates a symmetrical linear triangular waveform with a peak-to-peak amplitude of 0.33 times the supply voltage on TRIANGLE OUTPUT output pin 3, and a squarewave with a peak-to-peak value of the supply voltage on SQUAREWAVE OUTPUT pin 9, which is loaded by load resistor R_L .

The triangle wave is fed to the IC's internal sine converter which produces a good sine-wave output. Those sinewaves will have a peak amplitude of $0.22 \times$ supply voltage at SINEWAVE OUTPUT pin 2 in Fig. 5

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WAVEFORM GENERATORS

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when R_T has a value of 82 kilohms, as shown.

Sinewave distortion

The circuit in Fig. 5 will generate nonsymmetrical waveforms if the values of R_A and R_B differ. To obtain the purest sine waves, the circuit must be set to provide waveform symmetry. Figure 6 is a modification of Fig. 5 showing how fixed trimming resistor R_T is replaced by a trimmer potentiometer so that the sinewaves can be adjusted for minimum distortion.

This distortion of this circuit can be as low as 0.8% when it is set to produce fixed frequencies less than 10 kHz. Distortion can be further reduced by modifying the Fig. 6 circuit as shown in Fig. 7. In this circuit, all three trimmer potentiometers, R_3 , R_4 , and R_8 , can be trimmed to obtain optimum performance.

The ICL8038 rarely produces perfectly symmetrical waveforms if organized for variable frequency output applications. Distortion can be 2% or more. It should also be pointed out that the IC tends to run warm. It draws a quiescent current of about 12 milliamperes at 20 volts.

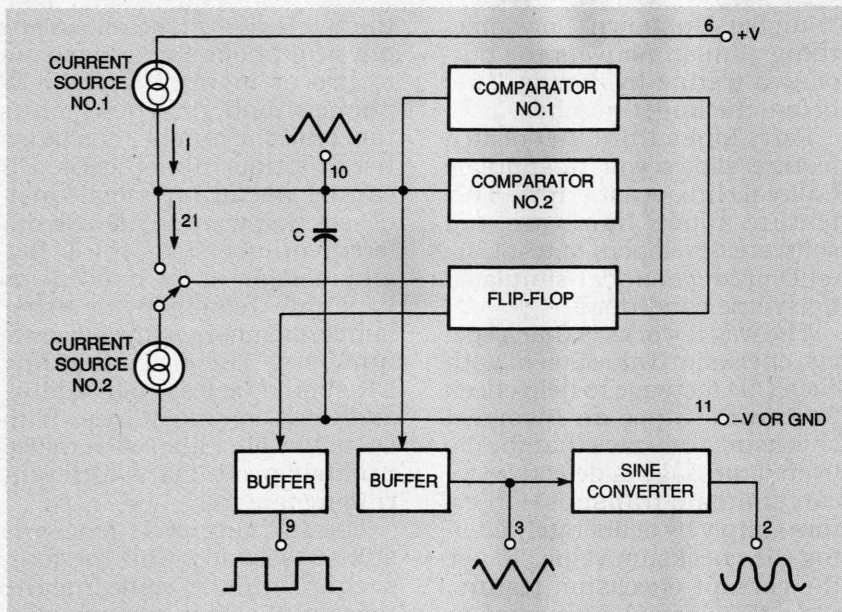


FIG. 3—A FUNCTIONAL DIAGRAM for the Harris ICL8038 waveform generator.

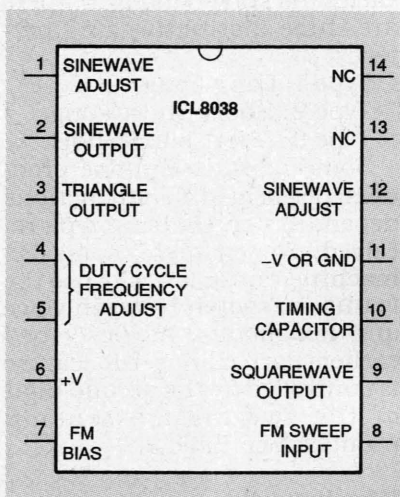


FIG. 4—PIN CONFIGURATION for the Harris ICL8038 waveform generator.

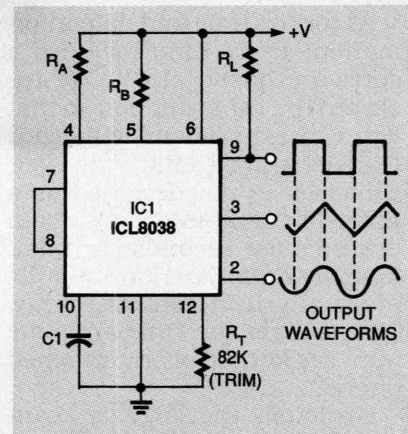


FIG. 5—FIXED-FREQUENCY triangle/sine/squarewave generator.

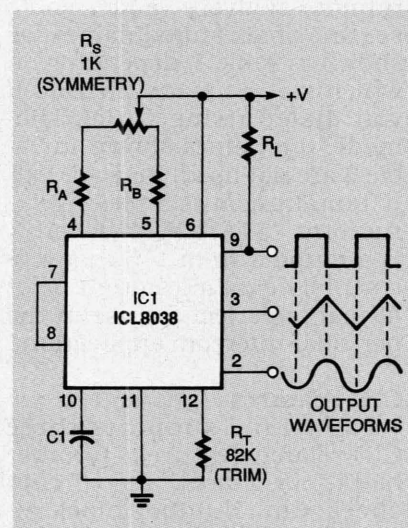


FIG. 6—MODIFIED FIXED-FREQUENCY triangle/sine/squarewave generator.

between pin 8 and the +V pin 6.

The operating frequency of the ICL8038 is directly related to the DC voltage applied between pins 6 and 8. As a result, frequency can be varied or swept by changing this voltage. The frequency can also be modulated by feeding a signal to pin 8. The circuit in Fig. 8, which includes an ICL8083, will function as a manually-controlled, variable-frequency waveform generator. Pin 8 is connected to the wiper of potentiometer R_1 so that the voltage can be varied from two-thirds of the supply voltage to its maximum value.

The lowest frequency is obtained when the voltage at pin 8 equals the supply voltage, and the highest frequency is obtained when the potentiometer wiper is set at two-thirds of the

Dual power supplies

The circuits shown in Figs. 5, 6, and 7 are all powered by positive or single-ended power supplies. The three output waveforms all swing about (or are centered on) the value equal to half the supply voltage. However, the circuits can all be powered by dual (positive and negative) power supplies. In those cases, the output waveforms will be centered on the ground bus of the dual supply.

Remember that regardless of the form of the power supply, FM SWEEP INPUT pin 8 is susceptible to unwanted signal noise and hum. The pin should be decoupled with a 0.1 μ F capacitor

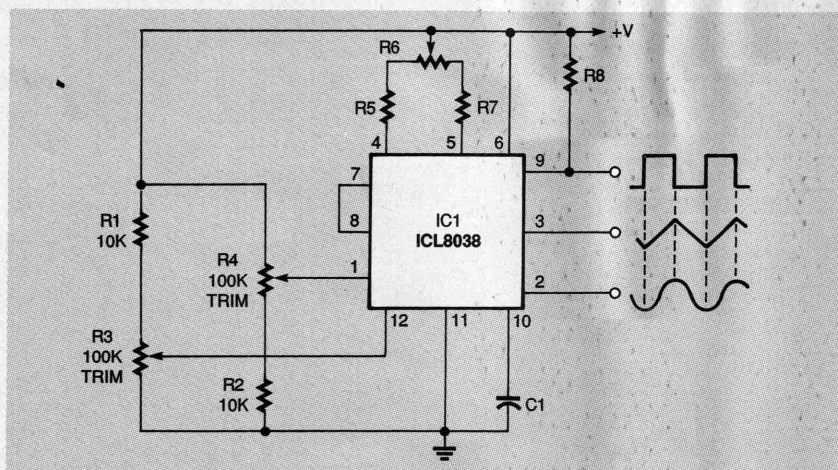


FIG. 7—SQUAREWAVE GENERATOR modified to minimize sinewave distortion.

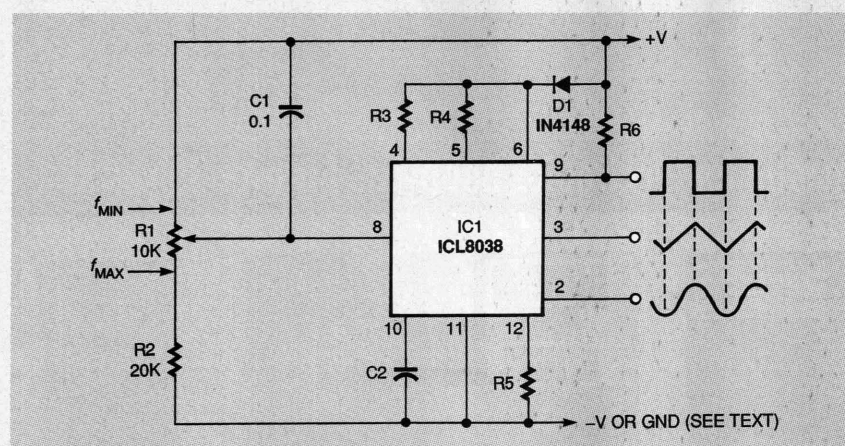


FIG. 8—WIDE-RANGE, VARIABLE-FREQUENCY waveform generator.

Table 1—Switch Position vs. Frequency

Switch 1 Channel	Frequency range (Nominal)
1	10 Hz to 100 Hz
2	100 Hz to 1 kHz
3	1 kHz to 10 kHz
4	10 kHz to 100 kHz

supply voltage plus 2 volts. This circuit can be varied in frequency over a range of about one thousand to one.

To obtain this range, the highest control voltage on pin 8 must exceed that on pin 6 by a few hundred millivolts. This is achieved by reducing the voltage on pin 6 to about 600 millivolts below the supply voltage with the forward drop of diode D1. However, for optimum frequency stability, both of the circuit's supply voltages must be regulated.

Figure 9 is a wideband (10 Hz

to 100 kHz) sine/triangle/squarewave generator formed by combining the circuits in Figs. 7 and 8 and adding an operational-amplifier buffer. To set up this circuit, set potentiometer R4 to its midrange and switch S1 to range 2. Then trim potentiometers R1 and R3 so that the generator can cover the 100-Hz to 10-kHz frequency range by adjusting potentiometer R2.

Next, set potentiometer R2 to give a 1-kHz output, and trim R4 to give a symmetrical squarewave output. Measure the frequency range, and then reset 1 kHz and trim potentiometers R9 and R12 for minimum sinewave distortion. Table 1 gives the frequency ranges that can be obtained from each of the four channel positions of switch S1.

Exar XR-2206

The XR-2206 is an integrated circuit function generator capable of producing stable and accurate sine, square, ramp and pulse waveforms. The output waveforms can be both amplitude and frequency modulated by an external voltage. They can also be frequency-shift

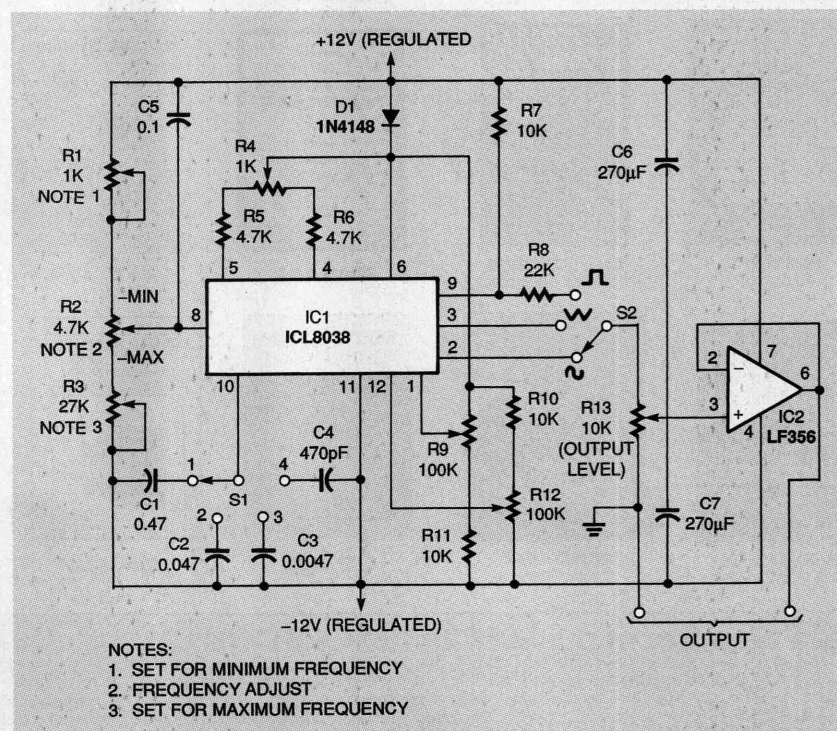


FIG. 9—WIDE-RANGE, VARIABLE-FREQUENCY waveform generator with a dual power supply.

key (FSK) modulated. The IC's operating frequency can be selected externally over a range of 0.01 Hz to more than 1 MHz.

Figure 10 is a functional diagram of the XR-2206 combined with its pinout arrangement diagram. The device includes four functional blocks: a voltage-controlled oscillator (VCO), an analog multiplier and sine shaper, and a set of current switches. It can be powered by either a single or dual power supply.

The VCO produces an output frequency that is proportional to an input current. That current flows in a resistor from the timing terminal to ground. The current switches route current from one of the timing pins to the VCO controlled by a frequency-shift keying (FSK) input pin to produce an output frequency. With two timing pins, two discrete output frequencies can be independently produced for FSK generation applications.

Figure 11 is a function generator circuit that includes an XR-2206 powered by a single power supply. Its operating frequency is inversely proportional to the values of C1 and the combined resistance of resistor R1 and potentiometer R7 in series. Frequency can be varied from 10 Hz to 100 kHz in four decade ranges with the capacitor values given in Table 2. The amplitude of the sine/triangle output can be varied with output-level potentiometer R8, but it can have its maximum value preset by level-control potentiometer R6. The sinewave distortion of this circuit is typically less than 2.5%.

Figure 12 is a modified version of the Fig. 11 circuit that operates from a dual power supply. It is designed to reduce sine-wave distortion to about 0.5 % by the adjustment of symmetry potentiometer R3 and distortion-control potentiometer R5. These two controls must be adjusted in unison after the circuit is built, but they need not be adjusted again. The maximum output level of these circuits can be preset by potentiometer R1. It should be set to

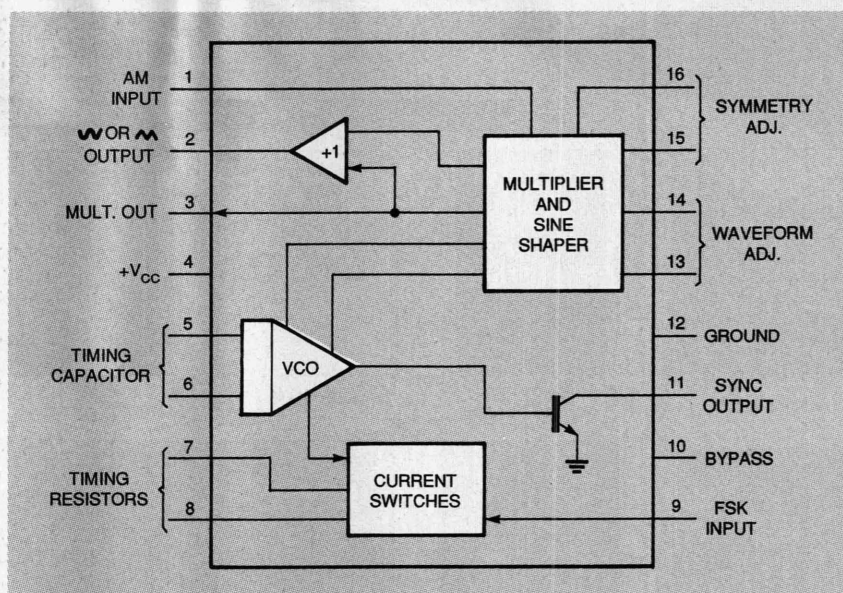


FIG. 10—FUNCTIONAL BLOCK DIAGRAM for the Exar XR-2206 sine/triangle/square-wave generator.

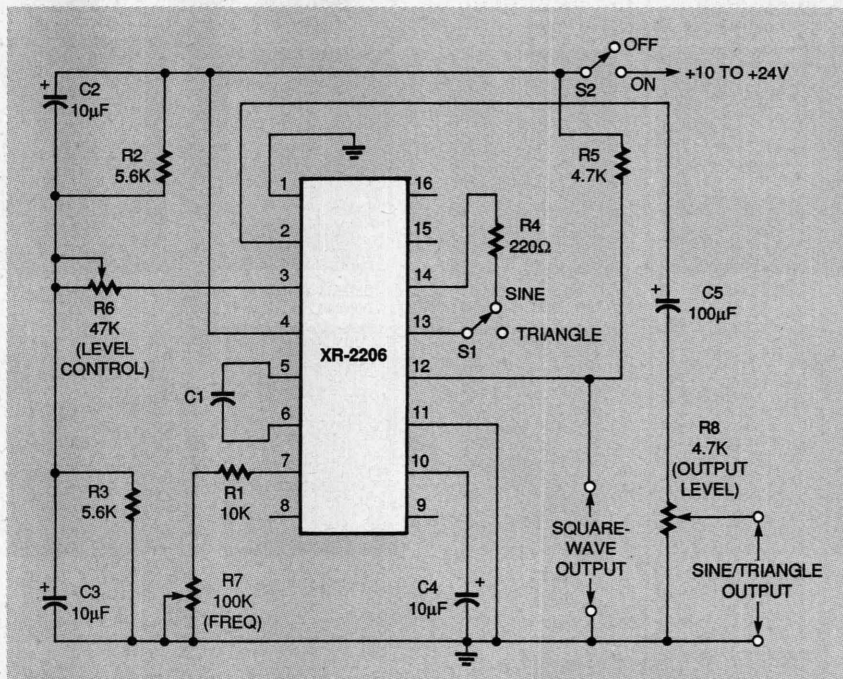


FIG. 11—SINE/TRIANGLE/SQUAREWAVE generator based on the XR-2206.

Table 2—Capacitor Values vs. Frequency

Capacitor C1	Frequency range (Nominal)
1 µF	10 Hz to 100 Hz
0.1 µF	100 Hz to 1 kHz
0.01 µF	1 kHz to 10 kHz
0.001 µF	10 kHz to 100 kHz

give a maximum output of less than 2 volts RMS to prevent excessive distortion. Table 2 lists

the frequency range of this circuit for each value of capacitor C1, as in Fig. 11.

LC oscillators

The last article in this series discussed the Wien-bridge and twin-T oscillators. Both included resistive-capacitive (RC) tuning networks whose output was limited to several hundred kHz. By contrast, oscillators that include inductive-capacitive (LC) tuning circuits can generate signals from tens of

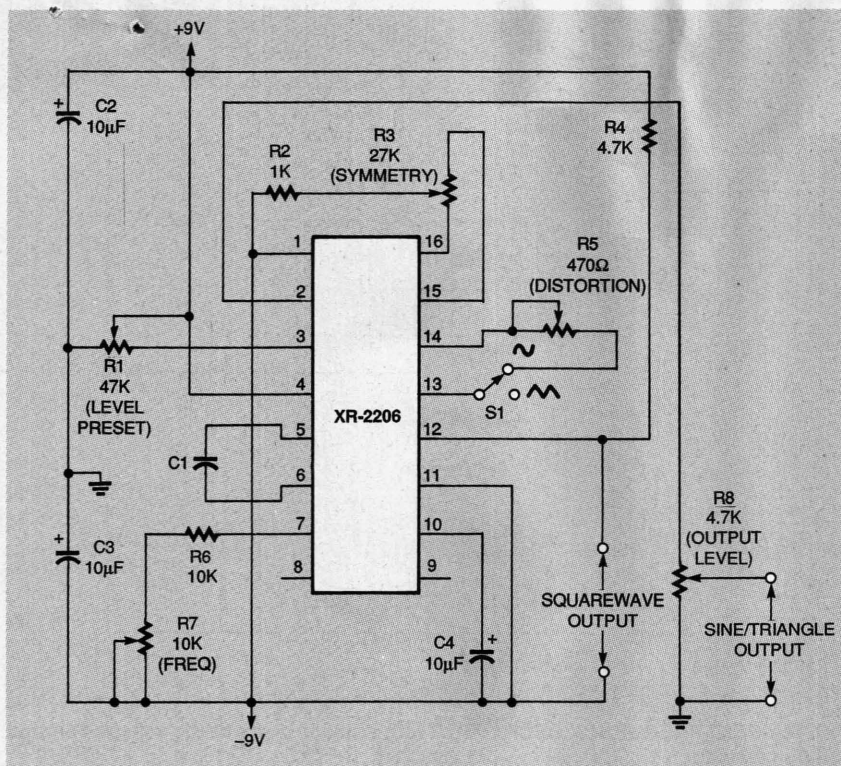


FIG. 12—SINE/TRIANGLE/SQUAREWAVE generator based on the XR-2206 with a dual power supply.

kHz to hundreds of MHz.

A transistorized LC oscillator includes a transistor amplifier and a frequency-selective LC network that provides feedback in accordance with the Barkhausen criteria. Inductive-capacitive networks have inherently high Q or frequency selectivity. Oscillators in this class produce reasonably pure sinewaves, even if the oscillator's loop gain is far greater than unity.

Among the many LC transistor oscillators are the tuned-collector, Hartley, Colpitts, Clapp, and Reinartz oscillators. Figure 13 is a schematic for a tuned-collector oscillator. In this schematic, the base bias for transistor amplifier Q1 is provided by resistors R1 and R2, and resistor R3 that is decoupled by capacitor C2.

The tuned-collector circuit is formed by L1-C1, and the collector-to-base feedback is provided by coil L2 of transformer T1, which is inductively coupled to coil L1. By selecting the phase of this feedback signal, the circuit can give zero-loop phase shift at the tuned frequency. Con-

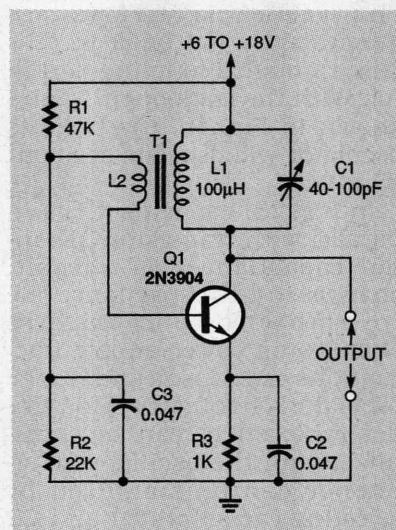


FIG. 13—TUNED-COLLECTOR inductive-capacitive oscillator.

sequently, if loop gain (determined by the turns ratio of T1) is greater than unity, the circuit will oscillate.

The phase relationship between the energizing current and induced voltage of all LC oscillators varies between -90° and $+90^\circ$. It is zero at the center frequency (f) determined by $f = 1/2\pi\sqrt{LC}$. The overall phase shift of the circuit in Fig. 13 is zero,

and the circuit oscillates at this center frequency.

If a tuned-collector oscillator has the component values given in Fig. 13, its center frequency can be varied from 1 MHz to 2 MHz by variable capacitor C1. However, the circuit can easily be modified to operate at frequencies ranging from tens of Hz (with a laminated-iron-core transformer) up to hundreds of MHz.

Hartley, Colpitts, and Clapp

The best-known LC oscillators are the Hartley and Colpitts oscillators, widely used in radio-frequency applications. The Hartley oscillator, as shown in Fig. 14, is a variation of the tuned-collector oscillator shown in Fig. 13. Its frequency-determining resonant tank contains a tapped coil. The collector and base are at the opposite ends of the tuned circuit to provide a 180° phase shift. Feedback occurs through mutual inductance between the two parts of the coil.

In Fig 14, the collector load coil L1 is tapped at a point 20 % along its overall length, and that tap is connected to the oscillator's positive power supply. Inductor L1 acts as an autotransformer, so the signal voltage that appears at the top of L1 is 180° out-of-phase with the signal voltage at its lower (collector Q1) end. The phase of the

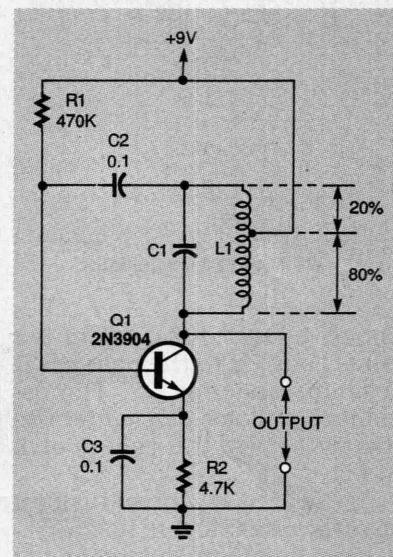


FIG. 14—HARTLEY TUNED-COIL inductive-capacitive oscillator.

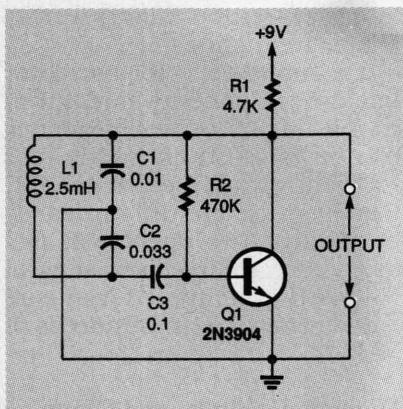


FIG. 15—COLPITTS TUNED-CAPACITOR oscillator.

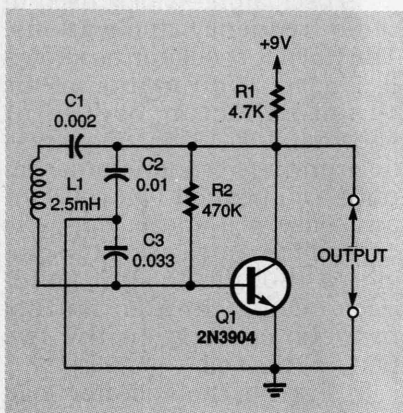


FIG. 16—CLAPP (MODIFIED COLPITTS) LC oscillator.

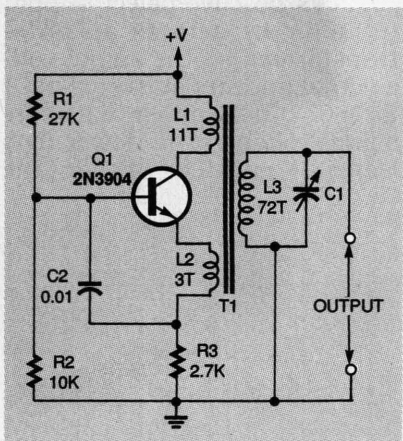
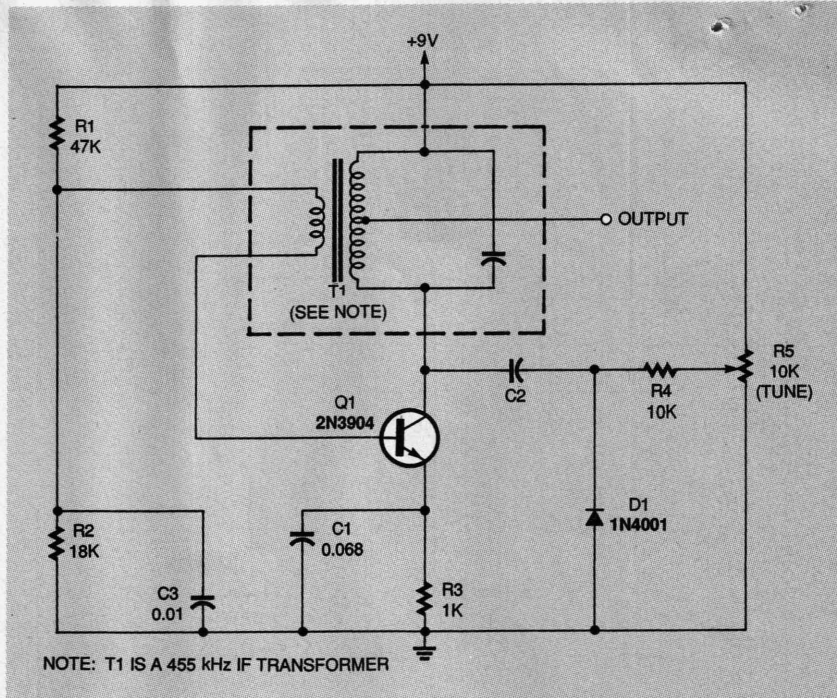


FIG. 17—REINARTZ LC oscillator.

signal voltage at the top of the coil is coupled to the base of Q1 through capacitor C2. The oscillator oscillates at a center frequency set by the values of L and C.

Figure 15 is a schematic for a Colpitts oscillator. It has a frequency-determining resonant tank circuit that includes a capacitive voltage divider. A frac-



NOTE: T1 IS A 455 kHz IF TRANSFORMER

FIG. 18—BEAT-FREQUENCY OSCILLATOR with a silicon diode as the voltage-variable tuning capacitor.

tion of the current flowing in the tank circuit is regeneratively fed back to Q1's base through a coupling capacitor. With the component values shown in Fig. 15, the Colpitts oscillator will oscillate at about 37 kHz.

In Fig. 15, capacitor C1 is in parallel with transistor Q1's input capacitance. As a result, changes in Q1's capacitance due to ambient temperature changes or power supply fluctuations can cause a shift in the oscillator's frequency. This undesirable effect can be minimized and the oscillator's frequency can be stabilized by selecting capacitors C1 and C2 with values that are large with respect to transistor Q1's internal capacitance.

Figure 16 is a schematic for a Clapp oscillator, a modification of the Colpitts oscillator. Its resonant circuit contains a small-value capacitor added in series with the resonant tank coil. The values of that capacitor and the tank coil determine the oscillator's frequency. This modification minimizes the effects of amplifier capacitance on the output frequency.

In this circuit, capacitor C3 is

in series with L1, and its value is small with respect to the values of capacitors C1 and C2. Consequently, the circuit's resonant frequency is determined principally by the L1 and C3 values, and it is not disturbed by variations in the capacitance of transistor Q1. A Clapp oscillator circuit with the component values shown in Fig. 16 will oscillate at about 80 kHz.

Reinartz and cap tuning

Figure 17 is the schematic for a Reinartz oscillator. Regenerative feedback occurs between the collector and emitter of a common-base amplifier transistor Q1. Its frequency is determined by the values of the LC resonant tank circuit, which is inductively coupled to the collector and emitter coils.

The Reinartz oscillator oscillates at a frequency set by the turns ratio of the primary and secondary coils of transformer T1. Figure 17 shows typical transformer turns ratios for oscillation at several hundred kHz. Reinartz oscillators can generate sinewave frequencies up to the UHF region.

Figure 18 is a schematic *continued from page 149*

Lamps and lighting efficiency

Plus a new RGB-to-NTSC converter, ultrafast computers, and product development concepts.

JUST GOT A HELPLINE CALL FROM AN "INVENTOR" TRYING TO "PROTECT" A "NEW" AUTO HEADLIGHT IDEA. TO STOP "DETROIT" FROM STEALING IT. I'VE NEVER HEARD OF "DETROIT" PAYING ANY OUTSIDER FOR AN

untested, undeveloped, or unproved idea. Instead, "Detroit" buys parts from suppliers, and bolts them together to make cars. Automobile manufacturers are in the process of outsourcing much of their product engineering. They are significantly *reducing* their number of suppliers, and holding them to the tightest of razor thin margins.

That's strike one.

Illumination engineering is one of the few things that Fortune 500 companies happen to do very well because it requires a multi-skill project team, combined with ray tracing computers, arcane production engineering, and outstanding access to the world's research base. So, those big boys clearly have an unbeatable home-turf advantage here.

That's strike two.

The most important requirement for any headlight advance is *efficiency*. It is becoming more important because of downsizing in general and electric or hybrid cars in particular. Anything less than 100 lumens per watt won't hack it. You can bet that tomorrow's headlights will *not* be based on a heated filament.

I got the impression the caller

was not a member of the SAE or the IESNA. Nor did he seem to be familiar with the relevant trade journals or to use online sources for research. He seemed to feel that car headlamp efficiency was "not important," and he apparently didn't have the slightest idea about

how woefully inefficient his new design was.

That's a swing and a miss for strike three.

A realistic alternative

Let's assume that you feel that you do have a great "new" headlamp design. What can you do that works out in the real world? Step one is to immerse yourself in relevant trade publications, industry associations, and online information sources so you can make sure that you weren't talking about a product that's been sold for

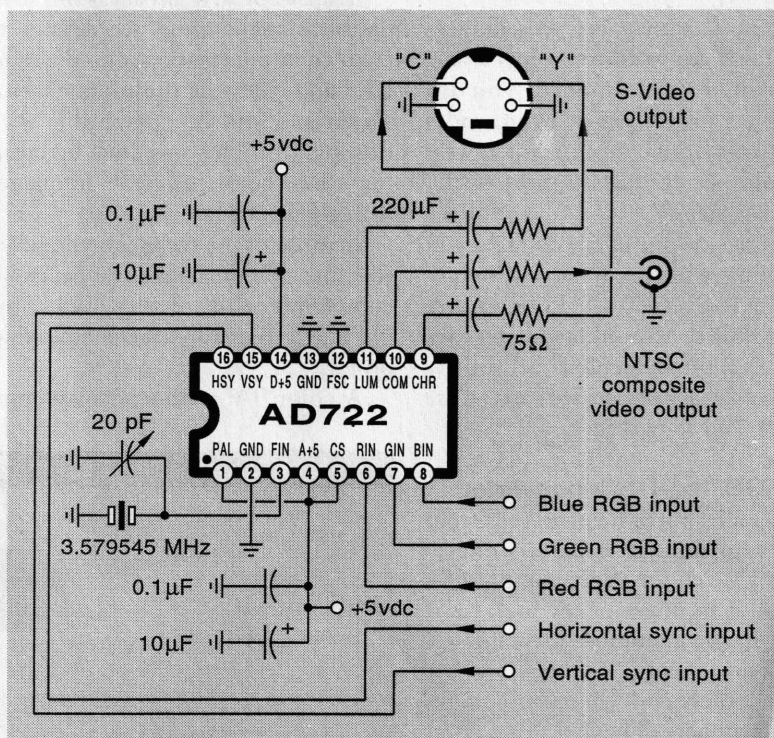


FIG 1—A NEW \$8 RGB-TO-NTSC CONVERTER CIRCUIT. Note that the input sweep rates, interlace, overscan, and program content must be TV-compatible if you want useful results.

decades—or one that long ago fell off the shelf because of inherent problems.

Step two is to ask yourself, “Who likes bright headlights, and have their own wallets in their own back pockets?”

Well, out here on my sand dune, the answer is glaringly obvious: four-wheel-drive desert off-roaders. To these folks, a map light is 50,000 candlepower, and a “running” light can vaporize troublesome boulders at 75 paces—on low beam.

So, first, you would have a few four-wheelers critique your design. If it is any good, you then let the local 4WD club beta test it. Once you have your tested and proven product well received, you sell a few at regional meets. Then you promote it in all the offroad magazines.

Next you seek out one or more of those off road lighting outfits. *K.C. Manufacturing* is but one of the name brand biggies out there. Competitors include *Dick Cepek*, *Hella*, *Explorer*, and *Piaa* corp. But be sure to remember the key insider secret rule for *all* successful new product development: *They must come to you.* And *never* vice versa.

Do note that you are *not* selling an idea. Ideas are worth ten cents a bale in ten-bale lots. Instead, you are offering a proven, in-demand, and a ready-to-manufacture product. You have already completed most of the high risk steps.

There is more of my thoughts on becoming a purveyor of risk reduction in *RISKDOWN.PDF* on my *GENIE* PSRT. More thoughts on idea development in general are in my *Blatant Opportunist* and my *Case Against Patents* packages.

An RGB to NTSC converter

I get all sorts of helpline calls from folks who want to use an ordinary color TV set for their computer—instead of buying a pricey high resolution RGB monitor—preferably via the antenna input.

Well, these days, you just can't get there from here. The TV's resolution is too low and the computer screen information content is way too high. Why do you suppose computers are

equipped with all of those fancy high-resolution monitors in the first place?

No matter what you do, there is *no way* you that can display “ordinary” modern computer color spreadsheet or word processor output by RF entry on an unmodified television! All you will get is a hopelessly smeared and violently flickering illegible mess.

On the other hand, if you design your computer screen *content* to be consistent with what your TV set can reasonably and intelligently handle, then you certainly can display it. At least after format conversion. Thus, computers can be used for video titling, for *large* screen data or text, for video editing, for “slide show” presentations, or for abstract art effects.

Analog Devices has developed a greatly improved and easy to use AD722 single-chip RGB-to-NTSC encoder. I've shown a typical circuit in Fig. 1. You can use this IC to convert *already TV viewable* RGB computer output into signals a TV set can understand. But you *can not* use this (or any other scheme I know of) to let a TV replace a high-resolution monitor.

The NTSC TV broadcasting standard was conjured up decades ago to cram color information in to the same RF bandwidth as the monochrome signal and still be compatible with existing sets. This was done by creating a new color *subcarrier* frequency of 3.579545 MHz.

Because of the technology limits at the time, and the need to be backward compatible with existing TVs, the NTSC system required many compromises.

A color TV cannot display com-

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puter data because (A) its horizontal and vertical sync rates, are far too low, (B) it has a severely limited resolution that prevents more than a few dozen color changes on any screen line, (C) its interlaced display causes flicker that makes small characters unreadable, and (D) its overscan hides some of the display.

If you are able to resolve *all* of the problems mentioned, then building an RGB to NTSC converter makes sense. Otherwise, it does not.

Briefly reviewing, video on an RGB system is on separate red, blue and green lines. An extra line or two

when adjacent lines are fairly similar. Since this is *never* the case with small dot matrix characters, small text will flicker badly.

The NTSC signal is purposely overscanned, gruesomely so on older sets. The reason is that if the picture is wrapped around the side of the tube, there will be no possibility of seeing black stripes on the side. Thus *all* NTSC programming *must* guarantee that the useful stuff always ends up *only* in the center.

The chip is easy to use. You first apply the usual +5 volts at 30 milliamperes. Also connect pin one to

else put composite sync into pin 16 while making pin 15 positive.

If you have a 4× color subcarrier clock available, input it to pin 3. If not, you can hang a stock *parallel resonant* color burst crystal from pin 3 to ground. Note that resynchronizing to a local crystal may degrade your results.

A logic high on pin 12 selects the 43 subcarrier mode, a low picks the 13 mode.

Be sure to read the data sheet carefully. There are several nasty "gotchas" to getting this chip to interact with VGA and other standards.

Conventional TV receivers require a chroma *delay line* to compensate for color shifts; a circuit equivalent to a delay line is *included* on chip so the results end up nearly identical to a broadcast signal.

The outputs are all at *twice* normal amplitude, letting you directly drive a 75-ohm reverse-terminated load.

Most VGA sources have provision for control by software. Once again, your sync rates and picture content *must* be NTSC-compatible ahead of time, or you'll get useless results.

Lighting Fundamentals

If you pick up any plain old 60 watt light bulb, it is likely to be rated around 850 *lumens*. That sounds like a bunch, until you find out what a lumen *really* is. Then you discover how badly you've been ripped off.

Well, a *lumen* is a measure of the *total output light power*. How many watts of light you will get back after gathering together all the output from all directions. (Well, usually, anyway. Some narrow-beam spotlights cheat and are rated in "effective lumens" instead.)

Being a unit of power, there is a relationship between the output light *lumens* and input electrical *watts*. This relationship also depends on the eye's sensitivity to colors. The eye responds with the *luminosity curve* shown in figure two. Eye sensitivity is highest in the yellow-green range.

A lumen is defined as 0.001496 watts of yellow-green light. Or 668.5 lumens per watt. A 100% efficient 60 watt yellow-green light bulb can thus produce 40,110 lumens.

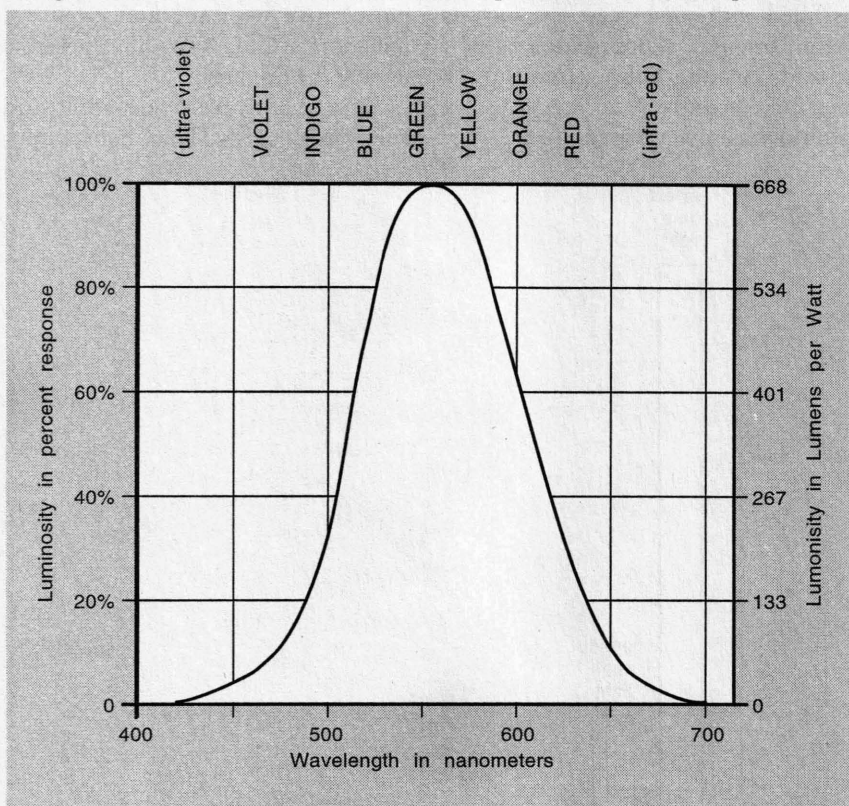


FIG 2—THE LUMINOSITY RESPONSE of the human eye varies with color. It is highest for yellow-green light, and lower for red and blue. White light sources are inherently less efficient than lime green ones.

is output for sync. Or else the sync gets combined onto your green channel. RGB scan rates are typically far higher than any TV can accept.

NTSC instead combines all of the picture information into a baseband black and white *luminance* signal and a color *chrominance* subcarrier.

The NTSC specification requires a pair of interlaced 262-1/2 line *fields* at a vertical sync rate of 59.96 Hertz. The horizontal scan rate is 15,735 Hertz. Interlace reduces flicker *only*

+5 to select NTSC instead of the PAL operation. The RGB outputs of your computer can be fed directly to the red, green, and blue inputs. Three outputs are provided: the composite video, a Y luminance output, and a C chrominance output. These latter two outputs can optionally get routed to an S-video connector.

In addition, you have to get sync and the stable color subcarrier from somewhere. You can input horizontal and vertical sync on separate pins. Or

Which means your plain old light bulb has a yellow-green efficiency of a mere 2.1%! When the other colors are considered, the total barely gets up to seven percent or so. A rating of *fourteen* lumens per watt, which is a result of nearly all the input energy being converted into useless heat.

One solution to efficiency improvement is to operate the lamp hotter. That works, but it also dramatically reduces lamp life. One workaround is the *halogen cycle*, once known as the "quartz-iodine" approach.

The result is still an incandescent lamp, but the "quartz" part means that the bulb temperature can be higher, and the "iodine" vapor part sets up a magic cycle that combines with any tungsten that is boiled off of the filament during hot times, forming a tungsten-iodide gas that is stable only when hot. The gas redeposits the tungsten right back onto the filament when it cools.

The result is a modest improvement in efficiency—Around 15% better for a 52-watt Halogen replacement lamp. The higher initial costs will eat into your energy savings, but the bulbs do last twice as long.

The larger Halogen lamps can produce up to 22 lumens per watt. Two big "gotchas" here: You must *never* touch a halogen bulb with your fingers. The oil from your skin causes a cold spot which can shatter the glass. And the lamps must *not* be continuously turned on and off. The halogen cycle only operates properly when the lamp remains on for hours at a time, with lots of cooling time between uses.

Improving efficiency

Actually, the correct term here is *efficacy*, since the output energy is in a different form from the input. There are many lamp technologies far more efficient than any incandescent. Figure three shows how all these alternates compare.

Light emitting diodes— Many new LEDs are actually *more* efficient than incandescent bulbs, and they last a lot longer, too. Cavers have long ago picked up on all of those orange superbright *Hewlett-Packard* diodes as

backup light sources. The automotive industry is very excited about LED tail lights, both because of their longer life and their "instant on" feature. The latter translates to some *twenty feet* of extra safety margin at thruway speeds.

There is no fundamental physical limit restricting LED efficiency. It is more economics, material science, visual coupling, total output, and a poor performance in the blue region of the spectrum that is holding things up.

Neon lamps— These are just a pair of electrodes in a glass enclosure that is filled with some inert gas such as neon (orange), xenon (blue), carbon dioxide (white), helium (purple), or mercury (green). A current causes a glow discharge and output light.

Neon lamps are potentially very efficient "cold light" sources. The lumens per watt depends on the color. Output power is normally low. High voltages are always required. More information on these is available in *Neon News*, *POP Design*, *Sign Craft* or *Sign Business* trade magazines.

Electroluminescent panels— These are another "cold light" source. Basically, an electroluminescent panel is a capacitor with a phosphor on one electrode. A high AC voltage (100-400 volts) is applied, and its field strength excites the phosphor. The panels can produce many colors, including white. A medium green is usually the brightest.

Best results are obtained at mid audio frequencies. Total light output

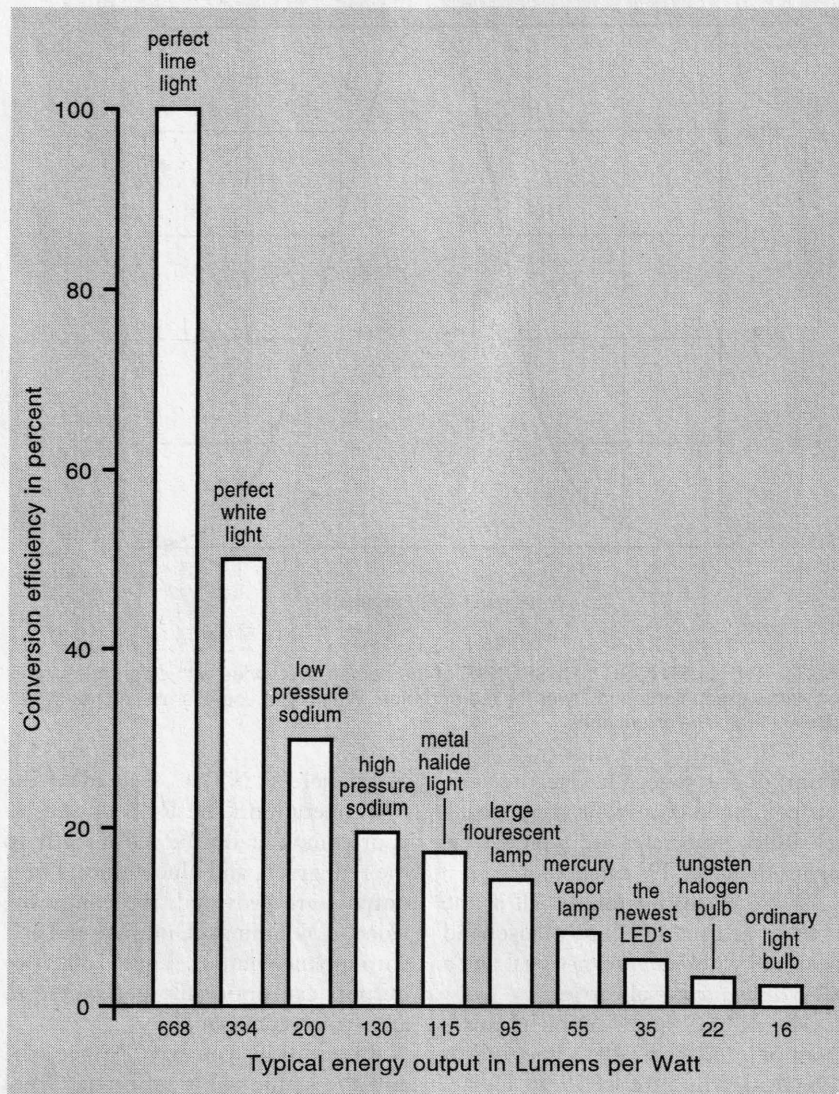
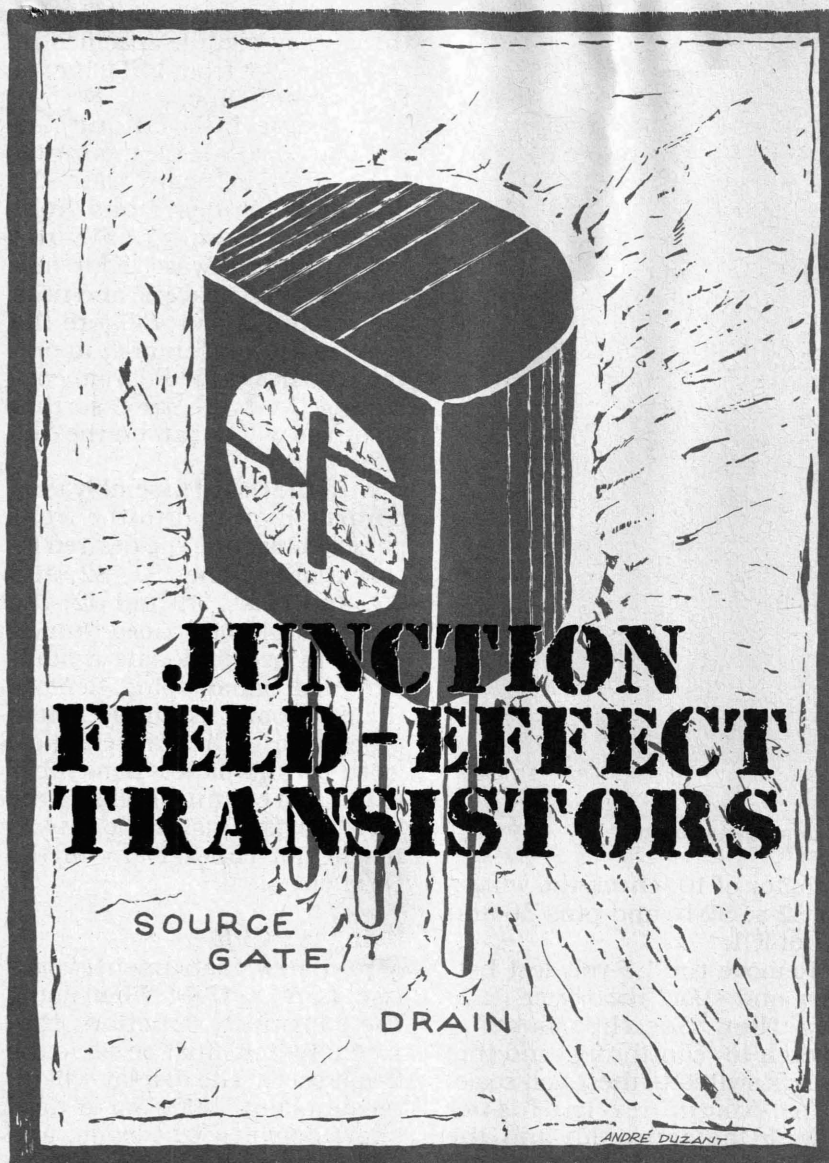


FIG 3—EFFICIENCY RATINGS for several emerging lamp technologies. Note that ordinary incandescent light bulbs are utterly awful.



JFET, has the same terminals as other FET's: *source*, *drain* and *gate*. The JFET is a unipolar device whose operation depends only on the movement of majority carriers—electrons or holes—not both as in bipolar transistors. As a voltage-operated transistor, the appropriate voltage applied to the JFET's gate controls the flow of current between the drain and source.

Figure 1-a is a cross-section view of a modern diffused planar N-channel JFET showing its three terminals and three doped regions: *substrate*, *source-to-drain channel*, and *gate*. Figure 1-b is the schematic symbol for an N-channel JFET. The vertical bar represents the normally conductive channel region. (Symmetry within the JFET permits the source and drain terminals to be interchanged.)

Figure 2-a illustrates the structure of a P-channel JFET. It is made the same way as the N-channel JFET shown in Fig. 1-a except that N- and P- doped regions are interchanged. The schematic symbol for the P-channel JFET, shown in Fig. 2-b, has its arrow pointed away from the bar representing the channel.

All JFET's operate in the *depletion* mode. This means that maximum current flows in the source-to-drain channel when the gate bias is zero. To reduce

Learn how to bias junction FET's and apply them in practical amplifier, voltmeter, multivibrator, and converter circuits

RAY MARSTON

THE JUNCTION FIELD-EFFECT TRANSISTOR (JFET) has interesting characteristics that set it apart from the bipolar transistor. This article contains schematics for a selection of practical JFET circuits that range from amplifiers and analog voltmeters to a multivibrator and a DC- to AC-converter. Last month's article discussed the differences between JFET's and MOSFET's, and FET terms were defined.

You might want to read—or reread—last month's article before you tackle this one unless you are really “up to speed” on FET's, their operation, symbols and terms. The many combinations and permutations of N- and P-channel FET's operating in *enhancement* and *depletion* modes can be very confusing even for the experienced circuit designer.

The JFET reviewed

The subject of this article, the small-signal, general-purpose

(*deplete*) or entirely *pinchoff* that current, the gate must be reverse biased. In an N-channel JFET, a negative bias must be applied, while in a P-channel JFET, a positive bias must be applied.

Figure 3 is a family of drain characteristic curves for an N-channel JFET. Note that the amplitude of the drain current (I_D) decreases as gate bias (V_{GS}) becomes more negative from V_{GS} equals zero. The family of curves for a P-channel JFET are similar except that the bias val-

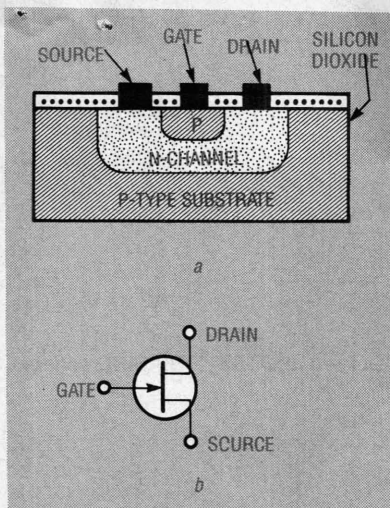


FIG. 1—AN N-CHANNEL DIFFUSED JFET (a), and schematic symbol (b).

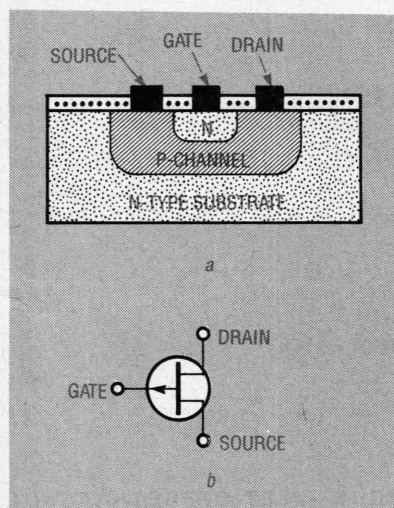


FIG. 2—A P-CHANNEL DIFFUSED JFET (a), and schematic symbol (b).

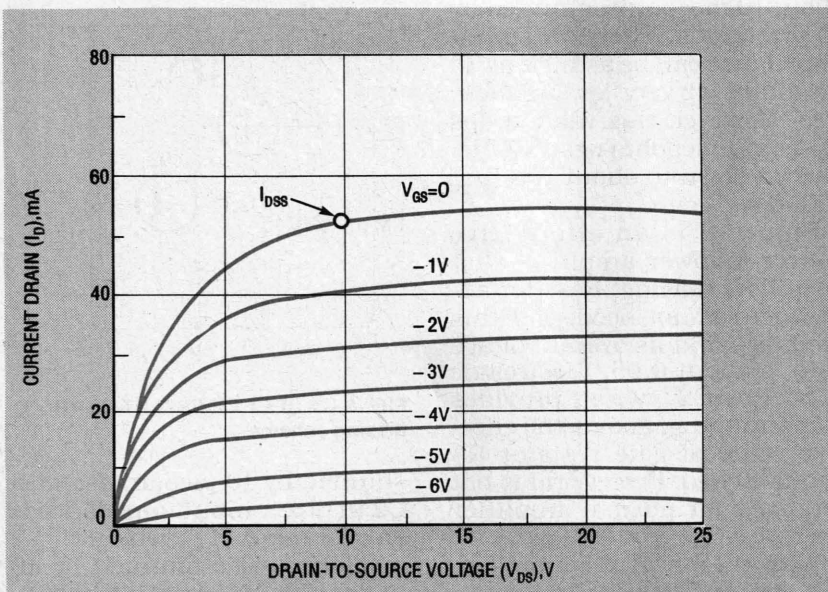


FIG. 3—N-CHANNEL JFET DRAIN characteristic curves.

ues become more positive with decreasing drain current.

All of the schematics in this article are based on the classic 2N3819 N-channel JFET. First introduced more than 25 years ago, it is packaged in three-pin TO-92 plastic case. Table 1 gives the maximum ratings for this device at 25°C free-air temperature. If you want to build these circuits with an N-channel JFET other than the 2N3819, be sure that the substitute's electrical characteristics are closely matched to those of the 2N3819. Also, be sure that the pinout arrangements are similar.

Biasing the JFET

The JFET will work in digital as well as linear circuits. In a low-distortion analog amplifier, it must operate in its linear region by reverse biasing its gate relative to its source. There are three common JFET biasing techniques: *self*-, *offset*-, and *constant-current*.

Self-biasing is shown in Fig. 4. The JFET's gate is grounded through resistor R_G , and resistor R_S grounds the source. Any current flowing in R_S drives the source positive with respect to the gate, so the gate is effectively reverse-biased. If drain current (I_D) is to be set at 1 milliampere, and it is known that a gate-to-source bias voltage (V_{GS}) of -2.2 volts is needed, the

correct value of source resistor (R_S) must be determined.

This correct bias can be obtained with a 2200-ohm value of resistor R_S . By Ohm's law, if 2.2 volts appears across a 2200-ohm source resistor, a 1 milliampere current will flow. If the drain current decreases, gate-to-source bias voltage also decreases. This causes drain current to increase and counter the original change. Thus, the bias is self-regulating through negative feedback.

The value of gate-source bias needed to set a desired drain current can vary widely even among identical JFET's in actual circuits. Thus, the only sure way to set a precise drain current is to pick a source resistor by trial and error or use a potentiometer. Regardless of how it is obtained, self-biasing is satisfactory for most practical applications, and only a few external components are needed. That's why it is still the most popular way to bias a JFET.

The second scheme, *offset biasing* is illustrated in Fig. 5-a. It gives more accurate gate biasing than self-biasing. Here, the voltage at the junction of resistors R_1 and R_2 is applied as a fixed positive bias to the gate through gate resistor R_G . The voltage at the source equals this bias voltage minus the negative value of the gate-source bias.

Therefore, if positive gate voltage is large with respect to gate-source bias, drain current is controlled mainly by R_S and gate-voltage; it is not greatly influenced by variations of gate-source bias between individual JFET's. Offset biasing permits drain current to be set accurately, avoiding the chore of individual resistor selection. Similar results can be obtained by grounding the gate and coupling the low-end of the source resistor to a high negative voltage, as shown in Fig. 5-b.

The third scheme, *constant-current biasing*, is illustrated in Fig. 6. The source resistor is replaced by NPN bipolar transistor Q2, which is organized as a constant-current generator. Consequently it determines the

TABLE 1
SILICON N-CHANNEL JFET: 2N3819
ABSOLUTE MAXIMUM RATINGS
(25°C Free air temperature)

Parameter			Unit
Gate-Source Breakdown Voltage	$V_{(BR)GSS}$	-40	V
Zero-Gate Voltage Drain Current	I_{DSS}	20	mA
Forward Transconductance	g_{fs}	7.0	mmho
Reverse Gate Leakage	I_{GSS}	-100	pA
"ON" Resistance	r_{DS}	500	ohms
Pinchoff Voltage	$V_{GS(OFF)}$	-6	V
Output Conductance	g_{os}	10	umho
Feedback Capacitance	C_{rss}	0.9	pF
Input Capacitance	C_{iss}	4.0	pF
Power Gain	G_{ps}	12	dB
Power Dissipation		360	mV

drain current. The constant current is set by Q2's base voltage, which is set from the R1-R2 voltage divider and emitter resistor R3.

Resistor R2 can also be replaced by a Zener diode or other voltage reference. Thus, in this bias circuit, drain current is independent of JFET characteristics, and high biasing stability is obtained. However, this improvement is gained at the expense of additional components.

In the three biasing schemes, resistor R_G can have any value up to about 10 megohms. That limit is imposed by the voltage drop across the resistor caused by gate leakage currents, which could upset biasing conditions.

Source-followers

JFET transistors in a linear amplifiers are usually configured as either a common-source or common-drain (*source-follower*) amplifier. These are the JFET equivalents of the bipolar common-emitter and common-collector (emitter-follower) amplifier, respectively.

The source-follower amplifier offers very high input impedance and near-unity overall voltage gain. (That's why it's also called a *voltage-follower*). A simple source-follower amplifier is illustrated in Fig. 7. It is self-biasing, and drain current can be varied with potentiometer R4.

That self-biasing source-follower amplifier will work from any positive 12- to 20-volt supply. Potentiometer R4 should be set so that the quiescent voltage

across R2 is 5.6 volts, which provides a 1 milliamperere drain current. Expect a voltage gain of about 0.95 between input and output.

Because of the voltage division at the junction of potentiometer R4 in series with R1 and resistor R2, some *bootstrapping* is applied to R3. In this circuit where the output is taken from the emitter, the output voltage directly affects the bias. In this amplifier, negative output pulses cause an increase in the negative voltage at the input, and positive output causes a reduction in the negative voltage at the input.

The input is applied between the source and the gate. In this circuit bootstrapping multiplies the effective value of R3 by a factor of about five. The input impedance to the circuit is about 10 megohms, shunted by 10 picofarads. Therefore, input impedance can be as high as 10 megohms at very low frequencies. However, this value drops to about 1 megohm near 16 kHz, and on down to about 100 K at 160 kHz.

Figure 8 is an alternative source-follower amplifier that has offset biasing. Resistor adjustment is not needed in this amplifier, and its overall voltage gain is about 0.95. Electrolytic capacitor C2, which provides bootstrapping, boosts the effective value of gate resistor R3 about 20 fold. However, it is not required for normal amplifier operation.

With C2 out of the amplifier, the source-follower's input impedance is about 2.2 megohms,

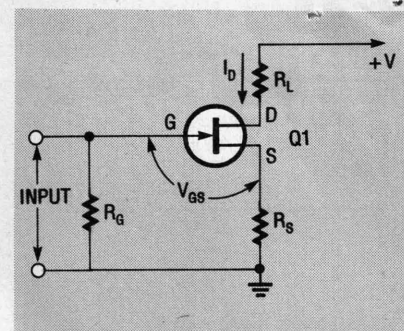


FIG. 4—A JFET SELF-BIASING scheme.

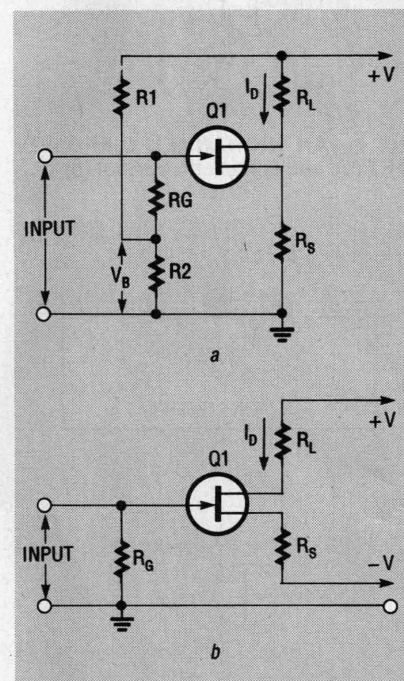


FIG. 5—JFET OFFSET-BIASING schemes for JFET's with +V supply (a) and +V and -V supplies (b).

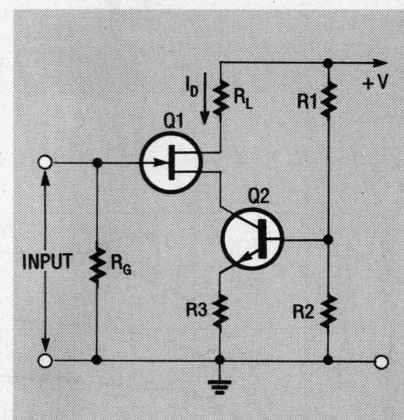


FIG. 6—A JFET CONSTANT-CURRENT biasing scheme

shunted by 10 picofarads; with C2 in place, input impedance is increased to about 44 megohms, also shunted by 10 picofarads. Alternative impedance values can be obtained by

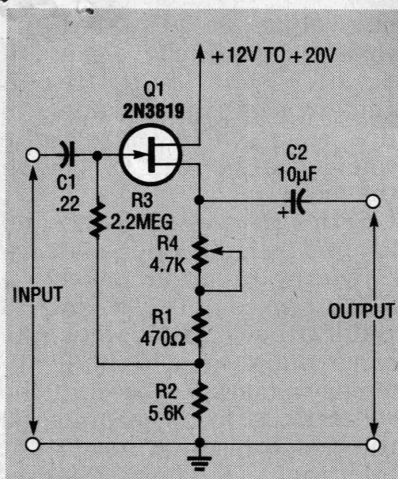


FIG. 7—A JFET SELF-BIASING source-follower with an input impedance of 10 megohms.

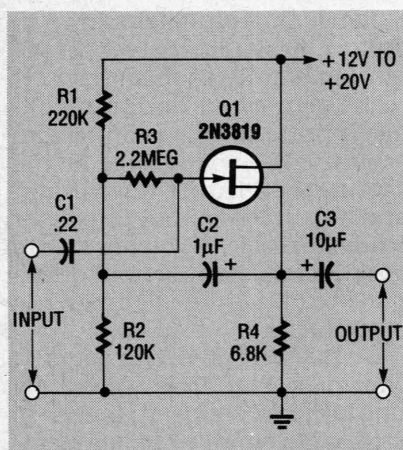


FIG. 8—THIS JFET SOURCE-follower with offset biasing has an input impedance of 44 megohms.

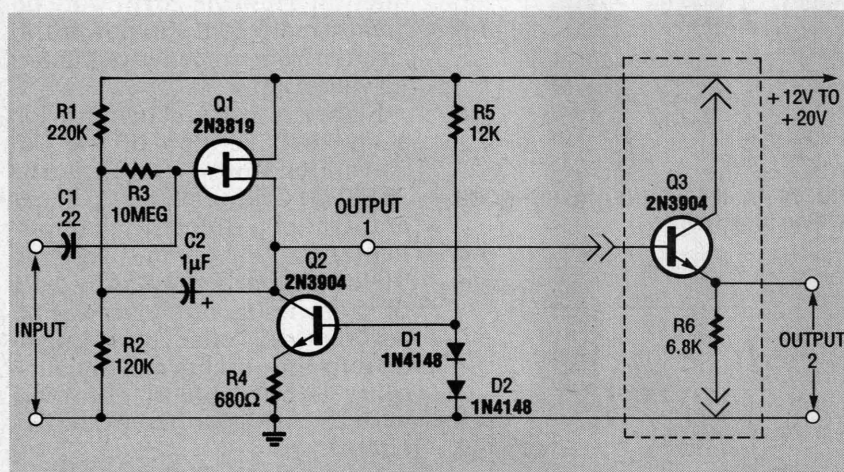


FIG. 9—THIS JFET SOURCE-FOLLOWER amplifier with bipolar transistors in its source circuit has an input impedance of 500 megohms.

increasing R3 up to a maximum of 10 megohms.

Figure 9 shows a JFET-bipolar "hybrid" version of a source-follower amplifier. Its in-

put impedance is about 500 megohms, shunted by 10 picofarads. Here, offset biasing is applied through a voltage divider formed at the resistor R1-R2 junction. This configuration is similar to that of Fig. 8, but source resistor R4 is replaced by a resistor-transistor-diode network (R4, Q2, D1, and D2).

This network makes "source load" Q2 act as a constant-current generator with a high output (collector) impedance, which causes a quiescent drain-to-source current of about 1 milliampere to flow in Q1.

As a result, Q1 functions as a source follower, and the collector of Q2, acting as its source load, appears as a high impedance. Because of the high effective value of this load, JFET Q1 has a voltage gain of about 0.99. Electrolytic capacitor C2 passes a bootstrap signal from the source of Q1 to R3 at the R1-R2 junction. The high-voltage gain of the circuit permits this bootstrap signal to boost the effective value of R3 100 times to about 1000 megohms.

As a result of bootstrapping, the actual input impedance of the source-follower in Fig. 9 is equal to 1000 megohms-shunted by the JFET's gate impedance of about 1000 megohms. The equivalent resistance turns

out to be about 500 megohms, shunted by 10 picofarads.

There are two ways to keep both the effective source load and input impedance of this cir-

cuit high: The output can be coupled to external circuitry with another emitter-follower stage (shown enclosed in dotted lines in Fig. 9), or all loads to which it is coupled must have high impedances.

Common-source amplifiers

Figure 10 is a schematic for a simple self-biasing, common-source amplifier that can be powered from any 12- to 20-volt supply. Potentiometer R4 should be adjusted so that a quiescent 5.6 volts is developed across R3, providing a drain current of 1 milliampere. The biasing of potentiometer R4 in series with resistor R2 is decoupled by electrolytic capacitor C2.

The typical voltage gain for the circuit shown in Fig. 10 is about 21 dB (a multiplying factor of 12), and its frequency response is flat within 3 dB from 15 Hz to 250 kHz. The input impedance of the circuit is 2.2 megohms, shunted by 50 picofarads. This comparatively high value of shunt capacitance is the result of Miller feedback from drain-to-gate. That feedback boosts the JFET's internal gate-to-drain capacitance directly with respect to voltage gain.

Voltage-biasing potentiometer R4 in Fig. 10 can be adjusted so that the circuit accepts, with minimal distortion, strong input signals that generate large output-voltage swings. In applications where only low-level input signals are to be accepted (such as in preamplifiers), a fixed-bias network can be substituted for the potentiometer. Figures 11 and 12 show circuits with that substitution.

Figure 11 shows a simple amplifier for headphones with impedances of 1 K or greater. With an input impedance of 2.2 megohms, the amplifier includes an integral volume control potentiometer R3, and it can be powered from any 9- to 18-volt positive supply.

The circuit shown in Fig. 12 is a general purpose, add-on preamplifier that can be coupled to any amplifier operating from a single-ended 9- to 18-volt

positive supply. The voltage gain from this preamplifier exceeds 20 dB, bandwidth exceeds 100 kHz, and its input impedance is 2.2 megohms.

When exceptional biasing accuracy is required, JFET common-source amplifiers can be designed with either of two *constant-current* offset biasing techniques. Figures 13 and 14 illustrate both options. The common-source amplifier in Fig. 13, with offset gate biasing, can only be powered by a 16- to 20-volt positive supply. However, the "hybrid" version shown in Fig. 14 can be powered with any 12- to 20-volt positive supply. Both circuits offer voltage gains of 21 dB, input impedances of 2.2 megohms, and -3 dB bandwidths from 15 Hz to 250 kHz.

DC voltmeters.

Figure 15 is a schematic for a simple three-range JFET analog voltmeter offering a nominal sensitivity of 22.2 megohms per volt. Its maximum full-scale voltage sensitivity is 0.5 volt, and its input resistance remains constant at 11.1 megohms on all ranges.

In the analog voltmeter shown in Fig. 15, resistor R6, potentiometer R9, and resistor R7 form a voltage divider across the 12-volt battery supply. With the proper setting of R9, 4 volts will appear across R7. The upper end of resistor R7 is connected to circuit ground (zero-volt reference), and its lower end is at -4 volts, setting the upper end of at +8 volts.

JFET Q1 is configured as a source-follower with its gate grounded through a resistor network consisting of R1 through R4. However, Q1's source is connected to -4 volts through source load resistor R5. As a result, Q1 is offset gate-biased, and its drain current is about 1 milliampere.

A closer look at the circuit shows that R6 in series with R9, and Q1 in series with R5 act as the arms of a Wheatstone bridge. The adjustment of potentiometer R9 balances the bridge, and no current flows in the meter unless there is an in-

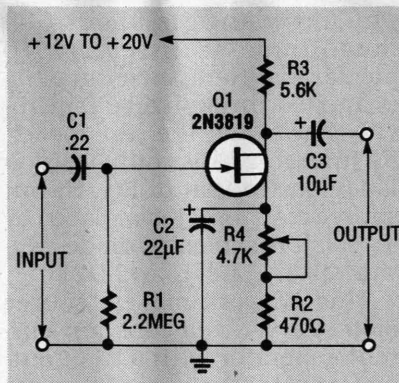


FIG. 10—A SELF-BIASING JFET common-source amplifier.

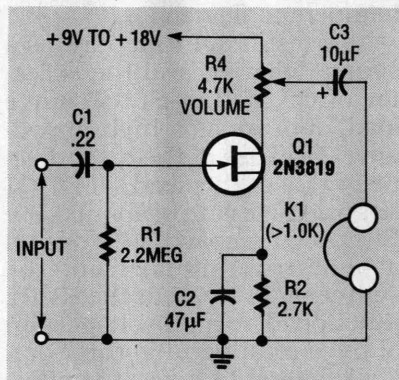


FIG. 11—A JFET HEADPHONE amplifier.

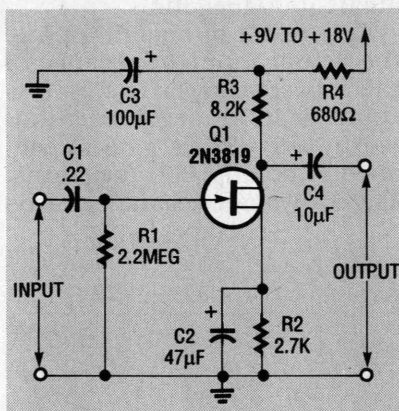


FIG. 12—A JFET GENERAL-PURPOSE add-on preamplifier.

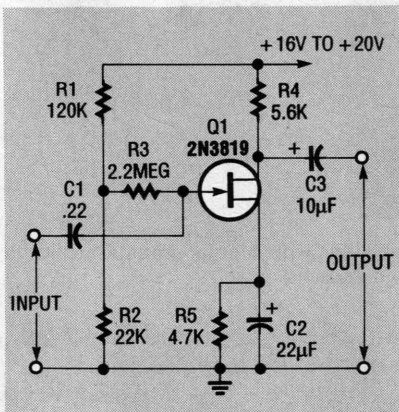


FIG. 13—A JFET COMMON-SOURCE amplifier with offset gate biasing.

put voltage at Q1's gate. Any voltage applied to Q1's gate unbalances the bridge by an amount that is proportional to that voltage. The value of the voltage can be read directly on the meter.

Series resistors R1, R2, and R3 form a multiplier network providing full-scale deflection (FSD) ranges of 0.5, 5, and 50 volts. Other resistor networks can be substituted to obtain different voltages. However, if the voltmeter is to be accurate, all the resistors must have tight tolerances. Resistor R4 prevents damage to transistor Q1 if the input voltage at the gate becomes excessive.

To use the voltmeter shown in Fig. 16, first adjust zero-set potentiometer R9 so that meter M1 reads zero when no input voltage is present. Then connect an accurate 0.5-volt source at the voltmeter's input terminals, and adjust R9 so that the meter reads full scale. If you are able to obtain consistent zero and full-scale readings, the voltmeter is ready for use.

Unfortunately, the circuit shown in Fig. 15 is inherently unstable because it tends to drift with changes in temperature and supply voltage. That makes it necessary to adjust zero-set potentiometer R9 frequently. However, drift can be greatly reduced if you power the circuit from a Zener-regulated 12-volt supply.

Figure 16 is the schematic for a low-drift version of the DC voltmeter shown in Fig. 15. JFET Q1 and resistor R5 form one arm of a differential amplifier, and JFET Q2 and resistor R6 form the other arm. Any drift that occurs in one arm will be automatically offset by a similar drift in the other arm. High stability is one benefit obtained with the bridge inherent in this circuit.

Ideally, Q1 and Q2 should be a pair of JFET's with their drain-to-source currents (I_{DS}) matched within 10%. This DC voltmeter circuit will work from any 12- to 18-volt positive supply. The calibration procedure is similar to that specified for Fig. 15.

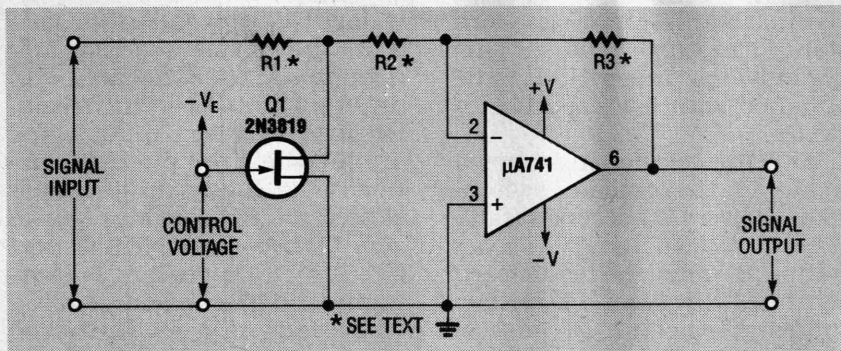


FIG. 18—A JFET CONTROLS AMPLIFICATION AND ATTENUATION for this op-amp.

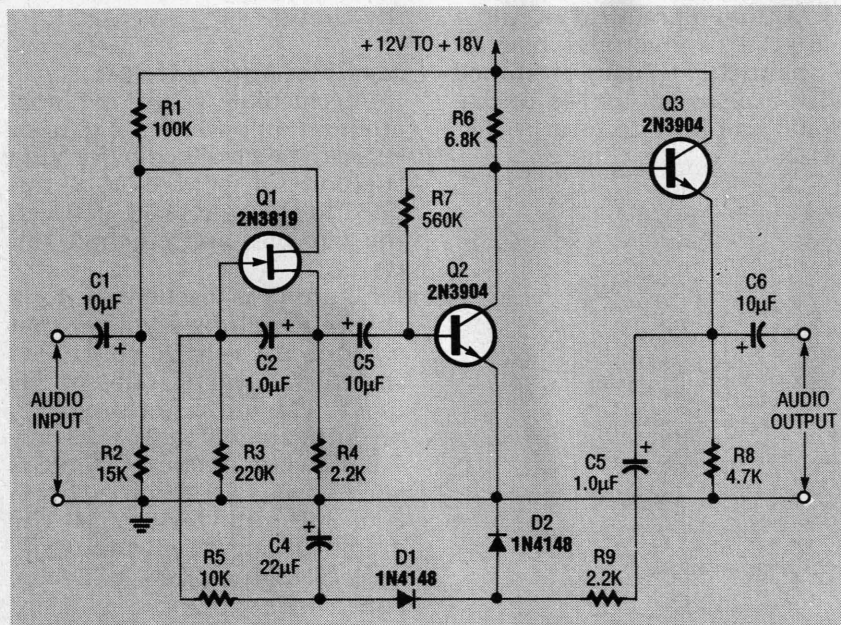


FIG. 19—THIS CONSTANT-VOLUME AMPLIFIER has a JFET in a closed loop for audio control.

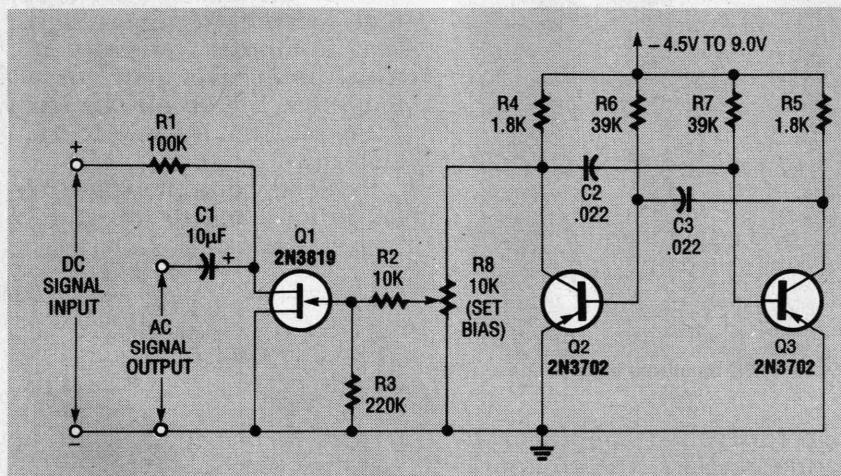


FIG. 20—THIS DC TO AC CONVERTER has a JFET switch and a bipolar multi-vibrator pulse generator.

JFET Q1 by the DC negative-feedback loop consisting of capacitor C5, resistor R9, diodes D1 and D2, capacitor C4, and resistor R5. The negative-feedback automatically adjusts the

constant-volume amplifier's voltage gain by holding the output signal level constant as the input signal level is varied.

When a small signal is applied to the audio input terminals of

the amplifier, the output at the emitter of Q3 is relatively small. Thus, very little negative bias is developed and fed to the gate of JFET Q1. JFET Q1 now acts like a low resistance and little signal attenuation occurs in the series resistance of the drop across Q1 and R4. As a result, most of the input signal is applied to the base of Q2.

However, when a large signal appears at the input terminals, the output at the emitter of Q3 is large. Thus, a large negative bias is developed and fed back to the gate of JFET Q1. In this condition, only a fraction of the input signal is applied to the base of Q2. Because of the negative DC feedback, the output level remains relatively constant over a wide range of inputs.

Figure 20 is the schematic for a DC-to-AC converter. The JFET acts as a switch, and a bipolar transistor flip-flop acts as the switching-signal generator. The converter produces a square-wave AC output with a peak amplitude equal to that of the DC input signal at its AC signal output terminals.

JFET Q1, the electronic switch, has its drain connected in series with input resistor R1 forming the positive DC terminal. JFET Q1 is switched on and off at a 1 kHz rate by the flip-flop consisting of bipolar transistors Q2 and Q3 and associated resistors and capacitors. The switching or "chopping" action converts DC to AC.

Potentiometer R8 varies the amplitude of the signal on Q1's gate. If Q1's gate signal is too large, its gate-to-source junction will avalanche, causing a voltage spike to appear at the drain. That transient will develop a small output signal at the drain of Q1 even when no DC input is present.

To prevent this glitch, connect a DC input to the converter, and adjust potentiometer R9 until the amplitude of the AC output just starts to decrease. JFET avalanching will not occur when this adjustment is made. As a result, the converter can reliably convert DC voltages with amplitudes as low as a fraction of a millivolt.

Q & A

READERS' QUESTIONS, EDITORS' ANSWERS

CONDUCTED BY MICHAEL A. COVINGTON, N4TMI

What Is Ground?

Q I have a beginner's question about the article you wrote, "Build the No-Parts PIC Programmer (NOPPP), which appeared in the September 1998 issue of *Electronics Now*, and about other project circuits. The diagram shows one common ground, but the picture of the NOPPP shows two pins both labeled "ground." Are they connected together? Are all these ground pins connected to the negative side of the batteries or power supplies? "Ground" is a confusing idea for me; I've read several basic books but none of them seem to explain it.—E. I., Durham, NC

A In my early days, "ground" puzzled me too. I knew that electrical power systems are grounded (connected to the earth) for safety reasons, and antenna systems sometimes use the earth as a counterpoise. But that didn't explain why almost all circuits have ground symbols in their schematics, even small battery-operated devices that can't possibly be connected to the earth.

Eventually I came across a book that defined "ground" as "common negative," or more generally, "common," and the light dawned.

Look at the diagram in Figure 1, which shows an imaginary but typical circuit (it could be some kind of DC amplifier). The diagram on the left uses the ground symbol; the diagram on the right does not.

The right-hand diagram shows that one side of the input, one side of the output, and one side of the battery are all connected together via the line at the bottom. This connection is "common to" (shared by) the input, output, and power supply. In the diagram on the left, the common connection is denoted by the ground symbol.

All the points marked by the ground symbol are connected together. Where the diagram shows only one connection for a two-terminal device, such as a battery, microphone, or speaker, the

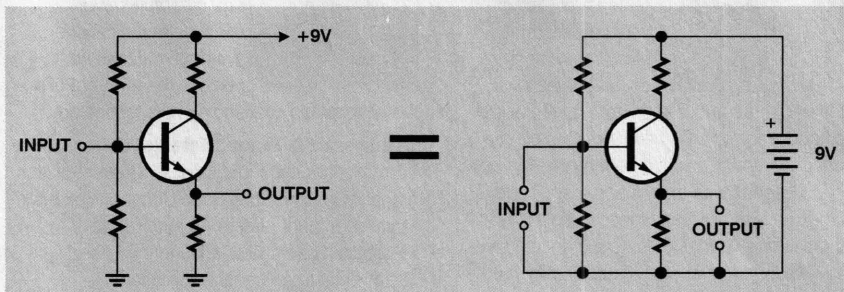


FIG. 1—A GROUND SYMBOL IN A SCHEMATIC shows points that are connected to each other and to the common side of the power supply, input, and output. As a result, these two simplified circuits are identical. Despite the term, ground does not imply a connection to the circuit's enclosure or to the Earth.

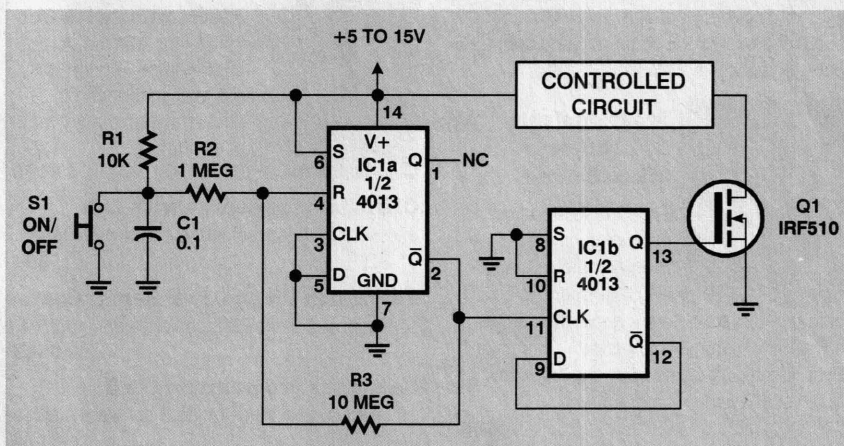


FIG. 2—HERE'S THE PUSH-ON, PUSH-OFF circuit that was originally published in the April 1998 installment of this column.

other terminal always goes to ground. Likewise, the "voltage at" any point in a circuit is measured by connecting a voltmeter between that point and ground. Cable shields are usually connected to ground.

Ground isn't always negative. A power supply marked -9V, as in an op-amp circuit, could consist of a battery with the negative terminal connected to that point and the positive terminal grounded. If the circuit shows +9V, -9V, and ground, it usually means that two batteries or the equivalent are involved, connected in series with the midpoint grounded.

Push-On, Push-Off, Again...

Q In the "Push On, Push-Off" circuit in your April 1998 column (and reprinted here as Fig. 2), the last paragraph states, "A disadvantage of the circuit is that you can't predict the state of the flip-flop when power is first applied."

I think you can force IC1b into the reset state with an RC delay circuit (see Fig. 3). While the capacitor charges, the reset pin is held high by the charging current; then the resistor holds the reset pin low so the flip-flop can function normally. When power is turned off, the diode rapidly discharges the capacitor to protect the IC, and the series resistor, R5,

have found one, or even a whole batch, that are "set in their ways" and happen to work consistently because of a mismatch between transistors on the chip.

As the diagram shows, the set and reset pins of the 74C74 are active low (not high) and are therefore inactivated by connecting them to V^+ rather than ground. A similar circuit with a 4013 would work (or not work) about as well, with the set and reset pins grounded.

... And Finally!

While researching these questions I came across a completely different way to build a push-on, push-off circuit (see Fig. 5). Two inverter gates make a Schmitt trigger that uses a capacitor for memory as well as debounce. The circuit always powers up with the load turned off, then switches on or off every time you press the button.

This is an adaptation of a circuit by T. J. Byers originally published in *Nuts and Volts*, May 1998. Here's how it works: Positive feedback ensures that the output of the second gate is always either fully "off" or fully "on." The output of the first gate is always in the opposite state.

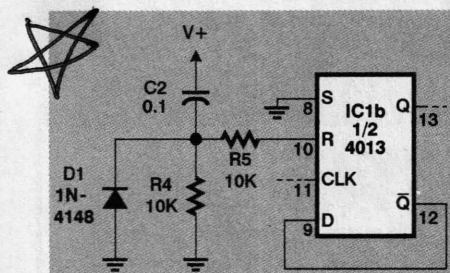


FIG. 3—THIS MODIFICATION TO APRIL'S circuit guarantees that the circuit is off when power is first applied.

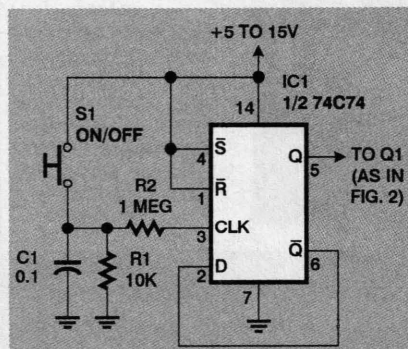


FIG. 4—WHILE UNDER SOME circumstances, the circuit shown here will work as a push-on, push-off switch, it relies on undocumented performance of the flip-flop. Also note that the flip-flop requires a fast-rising input signal.

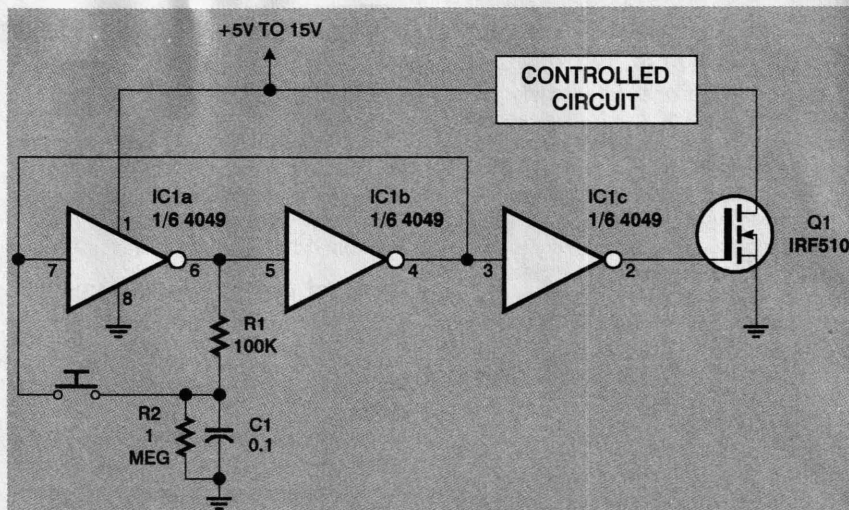


FIG. 5—HERE'S YET ANOTHER push-on, push-off switch. It draws less than one μA of current when in the "off state." Be sure to ground the unused inputs of IC1.

Resistor R2 ensures that the capacitor will be discharged if the circuit has been off for a while, so at power up, the output of the first gate is low. It stays that way until you press the button and connect the capacitor to the output of the second gate, which is high. Then the Schmitt trigger changes state; the output of the first gate goes high and keeps the capacitor charged. When you press the button again, the point to which you will be connecting the capacitor is now low, so the circuit will change state again. Debouncing is provided by the fact that the capacitor takes some time to charge. Unfortunately, the button is not returned to ground, so the buttons in some keypads are not usable.

While breadboarding this circuit, I learned another way *not* to use a digital multimeter (see Fig. 6). The power sup-

ply voltage seemed to be fluctuating strangely, in a staircase pattern, as the circuit switched on and off. Why? Because my autoranging digital multimeter, set on a low mA range, was changing its resistance as it automatically switched from one range to another. I turned off autoranging and everything was fine again. More tricks of modern technology!

Connecting Two Telephones Together

Q I have several ordinary telephones like you would find in a house. My daughters enjoy playing with them. Is there any way to connect the phones together so they can actually talk to each other without going through the phone line? Of course I would like them to ring when calling each other.—B. G., Clintwood, VA

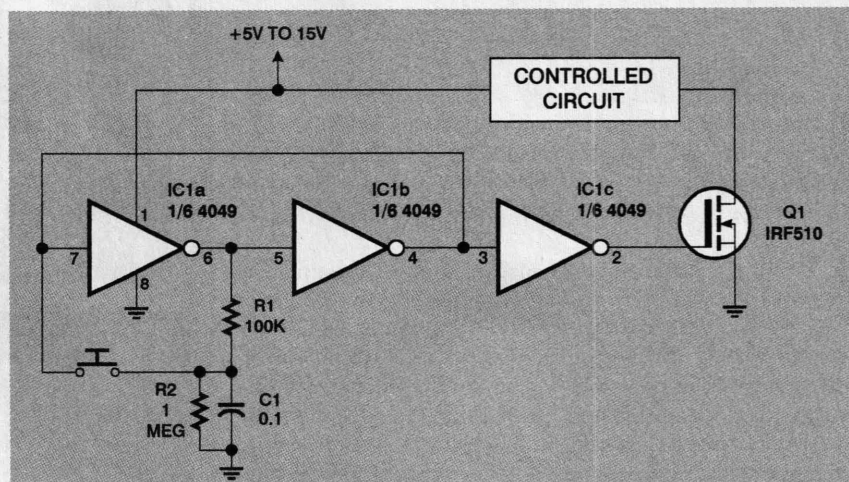


FIG. 6—AS THIS AUTHOR LEARNED from experience, avoid this situation when taking measurements with an autoranging meter. Otherwise, you'll get erratic results caused by changing meter resistance as the unit switches from range to range.

HOW TO GET INFORMATION ABOUT ELECTRONICS

On the Internet: See our Web site at <http://www.gernsback.com> for information and files relating to our magazines (**Electronics Now** and **Popular Electronics**) and links to other useful sites.

To discuss electronics with your fellow enthusiasts, visit the newsgroups sci.electronics.repair, sci.electronics.components, sci.electronics.design, and rec.radio.amateur.homebrew. "For sale" messages are permitted only in rec.radio.swap and misc.industry.electronics.marketplace.

Many electronic component manufacturers have Web pages; see the directory at <http://www.hitex.com/chipdir/>, or try addresses such as <http://www.ti.com> and <http://www.motorola.com> (substituting any company's name or abbreviation as appropriate). Many IC data sheets can be viewed online. Extensive information about how to repair consumer electronic devices and computers can be found at www.repairfaq.org.

Books: Several good introductory electronics books are available at RadioShack, including one on building power supplies.

An excellent general electronics textbook is *The Art of Electronics*, by Paul Horowitz and Winfield Hill, available from the publisher (Cambridge University Press, 1-800-872-7423) or on special order through any bookstore. Its 1125 pages are full of

information on how to build working circuits, with a minimum of mathematics.

Also indispensable is *The ARRL Handbook for Radio Amateurs*, comprising 1000 pages of theory, radio circuits, and ready-to-build projects, available from the American Radio Relay League, Newington, CT 06111, and from ham-radio equipment dealers.

Copies of past articles: Copies of past articles in **Electronics Now** and **Popular Electronics** (post 1994 only) are available from our Clagg, Inc., Reprint Department, P.O. Box 4099, Farmingdale, NY 11735; Tel: 516-293-3751.

Electronics Now and many other magazines are indexed in the *Reader's Guide to Periodical Literature*, available at your public library. Copies of articles in other magazines can be obtained through your public library's interlibrary loan service; expect to pay about 30 cents a page.

Service manuals: Manuals for radios, TVs, VCRs, audio equipment, and some computers are available from Howard W. Sams & Co., Indianapolis, IN 46214 (1-800-428-7267). The free Sams catalog also lists addresses of manufacturers and parts dealers. Even if an item isn't listed in the catalog, it pays to call Sams; they may have a schematic on file which they can copy for you.

Manuals for older test equipment and ham radio gear are available from Hi Manuals, PO Box 802, Council Bluffs, IA 51502, and Manuals Plus, PO Box 549, Tooele, UT 84074.

Replacement semiconductors: Replacement transistors, ICs, and other semiconductors, marketed by Philips ECG, NTE, and Thomson (SK), are available through most parts dealers (including RadioShack on special order). The ECG, NTE, and SK lines contain a few hundred parts that substitute for many thousands of others; a directory (supplied as a large book and on diskette) tells you which one to use. NTE numbers usually match ECG; SK numbers are different.

Remember that the "2S" in a Japanese type number is usually omitted; a transistor marked D945 is actually a 2SD945.

Hamfests (swap meets) and local organizations: These can be located by writing to the American Radio Relay League, Newington, CT 06111; (<http://www.arl.org>). A hamfest is an excellent place to pick up used test equipment, older parts, and other items at bargain prices, as well as to meet your fellow electronics enthusiasts—both amateur and professional.

limits the current applied to the reset pin under power supply transient conditions—Charles Hansen, Tinton Falls, NJ

A Thank you very much for a reliable, conservatively designed circuit modification. Not everyone would have thought of including the diode and R5, which aren't necessary under most conditions but are desirable to keep from overstressing the IC's input protection diodes and thereby make the chip last longer. We checked, and the 4013 data sheet doesn't specify any limits on rise or fall time of the reset signal, so there's no need for a Schmitt trigger or transistor in the reset circuit, as there would be with a microcontroller.

... And Again ...

Q I'm sending you a simpler push-on, push-off circuit that I've always had good luck with (see Fig. 4). I used a 74C74. The debouncing is done by R1 and C1. From my experience, the Q output is always 0 (low) at power-up.—A. M. DePaz, Globe, AZ

A Thanks for the circuit; it's a nice example of something that works better than it ought to (so to speak). According to the 74C74 data sheet, the flip-flop won't operate reliably unless the input signal makes the transition from low to high in less than 5 nanoseconds. In your circuit, the transition may or may not be that fast, depending on the impedance of the power supply. Fortunately, the flip-flop is clocked by the capacitor charging (which happens very fast), not discharging (a much slower process).

More importantly, there is no reason to expect the flip-flop to always power up in a particular state, though you may well



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WORKING WITH MIXERS

Mixer circuits are designed to operate as frequency translators. They are nonlinear circuits that produce an output frequency spectrum that includes various products of two input frequencies. As shown in Fig. 1, if two input frequencies, F_1 and F_2 , are combined in a nonlinear mixer, then the output spectrum will include the products:

$$F = mF_1 \pm nF_2 \quad (1)$$

where F represents the output products, F_1 and F_2 are the two input frequencies, and m and n are integers (0, 1, 2, 3...). In any given circuit the mixer products may be out to several higher orders. It is commonly, and erroneously, assumed that the output spectrum looks like Fig. 2. Assume that the two inputs are F_1 and F_2 . The outputs will include F_1 , F_2 , F_1+F_2 (sum frequency) and F_1-F_2

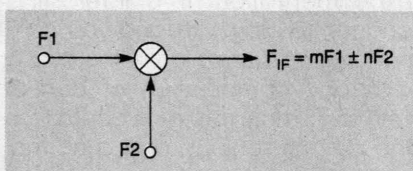


Fig. 1. If two input frequencies, F_1 and F_2 , are combined in a nonlinear mixer, then the output spectrum will include the products $F = mF_1 \pm nF_2$.

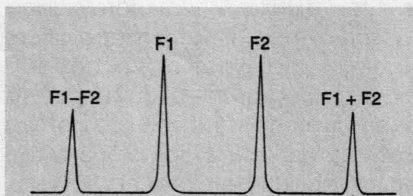


Fig. 2. It is commonly, and erroneously, assumed that the output spectrum of a mixer looks like the graph shown here.

As long as you follow some basic rules, mixer circuits are easy to understand and use.

JOSEPH J. CARR

(difference frequency). Of these frequencies, we usually only need one.

Although there are a number of applications for the mixer, let's take a look at only one. The most common application is the *superheterodyne* receiver design. Nearly all radio and TV receivers sold today are "superhets." In the superhet the incoming RF signal (F_{RF}) is converted to an *intermediate frequency* (IF) by "beating" a local oscillator (LO) signal (F_{LO}) against the RF frequency. That process is called *heterodyning*, from which the name *superheterodyne* was derived.

Figure 3 shows the block diagram of a typical superhet receiver. The radio signal at F_{RF} is picked up by antenna, and then (sometimes) amplified in an RF amplifier circuit. This signal is mixed with F_{LO} in the mixer to produce two output signals, in addition to F_1 and F_2 : F_1+F_2 (sum) and F_1-F_2 (difference). These frequencies correspond to the results of Eq.(1) when $m = n = 1$.

The IF amplifier is used to process the output signal of the mixer. Either the sum or the difference signal can be used for the IF, as they are mirror images of each other. In AM and FM broadcast band (BCB) receivers, the difference frequency is typically used. In the United States and

Canada, manufacturers usually use 455 kHz as the AM BCB IF, and 10.7 MHz as the FM BCB IF. In car radios, 262.5 kHz is used as the AM BCB IF. European and Japanese radios have sometimes used other frequencies for the IF, but they are close to these values. In recent years, we have seen high frequency (HF) shortwave receiver designs that use two conversion steps. The first IF is around 50 MHz, but this frequency is later down converted to a second IF such as 9 MHz, 8.83 MHz, or 10.7 MHz. In Fig. 3 the difference frequency $F_{LO}-F_{RF}$ is shown being selected.

The IF amplifier provides most of the signal gain found in the receiver. It also provides most of the selectivity of the receiver. The reasons for doing the frequency conversion is that the IF amplifier operates at one frequency only, and that means it is easier to obtain quality selectivity characteristics (the best filters are single-frequency devices) and high gain without either variation in gain with frequency or spurious oscillations.

Mixer Problems. One of the most common problems in using a mixer is the image response. Figure 4 shows how that works. The LO and RF are separated by the amount of the IF. Unfortunately, that means that there are two frequencies separated from the LO by the amount of the IF. In Fig. 4 the RF is located at $F_{LO} - F_{IF}$. Ideally, that is the only frequency that is received. But notice that $F_{LO} + F_{IF}$ (the "image" frequency) is also valid. Any signal that appears at the image frequency will also appear to the receiver as a valid IF signal.

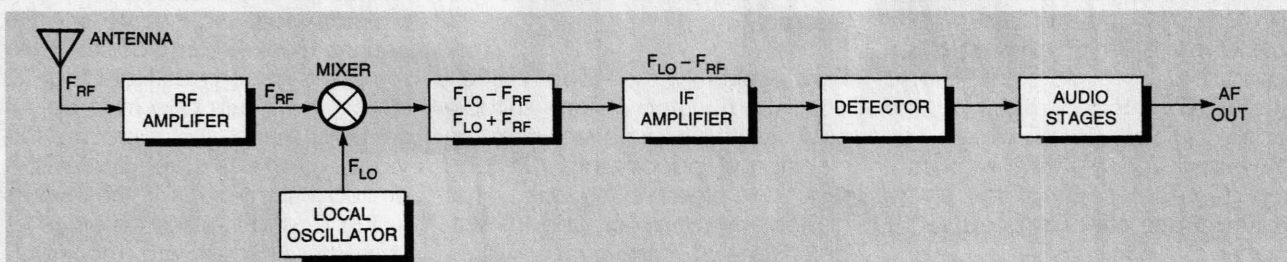


Fig. 3. A block diagram of a typical superheterodyne receiver. Though the difference-signal output of the mixer is used here, either the sum or difference signal output could be used.

The problem doesn't end there, unfortunately. The receiver has a certain bandwidth (BW). No signal has zero bandwidth, so a receiver designer will match the bandwidth with the modulation being received.

Typically, an AM BCB signal likes to see 10 kHz, an AM shortwave signal 6 kHz, an SSB signal 2.7 kHz, a land-mobile FM signal 5 kHz, and an FM BCB signal 200 kHz. Any signal that falls within the bandwidth will appear in the IF amplifier as a valid signal. Thus, all of the signals at $F_{RF} - (BW/2)$, i.e. around the RF signal in Fig. 4, and $F_{RF} + (BW/2)$, i.e. the image, will appear in the IF amplifier.

Notice in Fig. 4 that the baseline is not a nice clean line as it was in our ideal case of Fig. 2. Rather, there is a lot of noise in the system. The noise might be noise from the environ-

ment (both natural and man made), or it might be noise generated by the circuits inside the receiver. Even a simple resistor with no current flowing in it will produce a noise signal due to thermal agitation of the electrons inside the resistor material. Thus, in addition to any radio signals that might be present in the vicinity of the RF and image frequencies, any noise present will also appear in the IF amplifier.

It gets better: The local oscillator signal might be somewhat imperfect. Indeed, it will be imperfect... there's no "somewhat" about it. Three problems exist: LO harmonics, LO noise, and *discretes*.

Figure 5 shows the situation with respect to harmonics. All signals other than a perfect sinewave produce harmonics, i.e. signals of inte-

ger multiples of the basic signal. Thus, a 1000-kHz signal will have harmonics of 2000 kHz, 3000 kHz, 4000 kHz, and so forth, unless they are truly pure sinewaves (not likely).

In the example of Fig. 5 the LO signal (F_{LO}) has harmonics of $2F_{LO}$ and $3F_{LO}$. Any signals or noise that falls within the bandwidth (or pass-band) also appears around those harmonics plus or minus the IF: $2F_{LO} \pm (F_{IF} \pm (BW/2))$.

Figure 6 shows another problem with the LO signal: noise. The LO signal is ideally a spike with only amplitude and no width. That doesn't occur in practice. The real LO signal will have several different forms of noise around it. There will be signal components caused by thermal noise in the oscillator circuit. There will also be single-sideband phase noise present. These are difficult to filter out unless the LO is a single-frequency circuit, and its output signal can be passed through a very narrow (high Q) bandpass filter.

Discretes are modulation of the LO signal (either AM or FM) caused by other frequencies in the circuit. Of course, it's possible for oscillators and other signals within the circuit to modulate the LO, but the principal source of discretes is improperly filtered DC power-supply voltages. The AC power-line frequency in North America is 60 Hz (in some other parts of the world 50 Hz is used). When full-wave rectified, this produces a 120-Hz (twice the line frequency) ripple-frequency signal on the DC output of the power supply. If the ripple signal is allowed to remain on the DC power line, then it will modulate the LO signal and possibly cause problems.

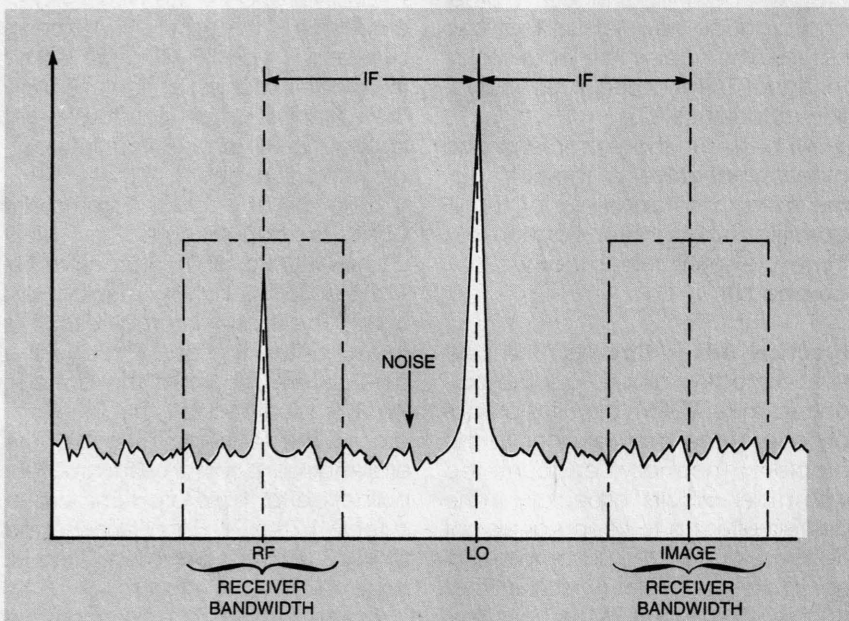


Fig. 4. While the RF signal located at $F_{LO} - F_{IF}$ is the desired one, an image signal located at $F_{LO} + F_{IF}$ is also valid.

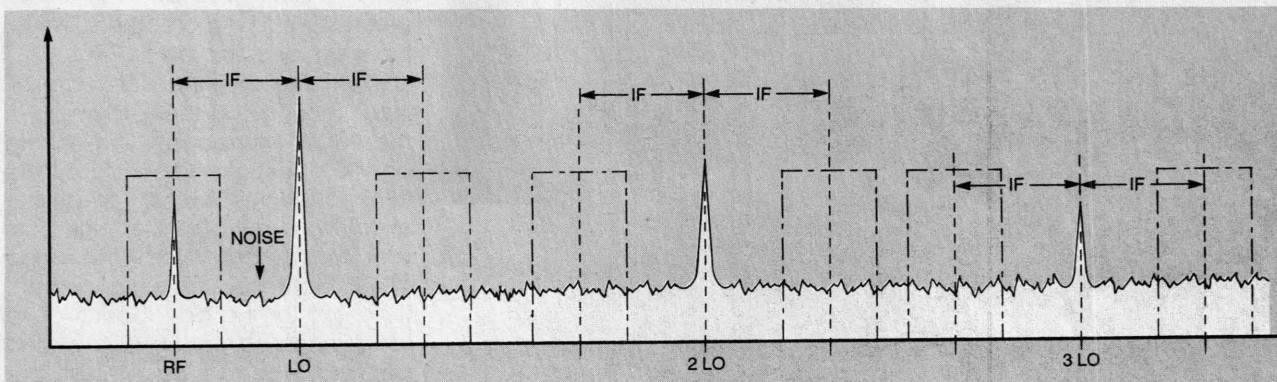


Fig. 5. Any noise that falls within the receiver's passband also appears around any internally generated harmonics.

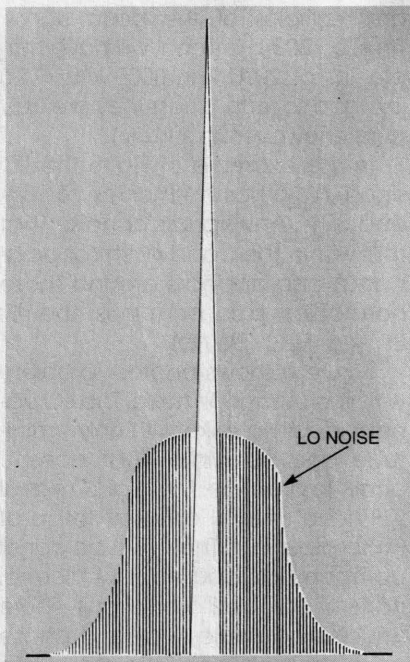


Fig. 6. While the LO signal is ideally a spike with only amplitude and no width, a real LO signal will have several different forms of noise around it.

Power supply discretely fall so close to the LO signal that they cannot usually be filtered out of the LO signal. They must be filtered out before they arrive at the LO. To cure this problem with the LO (as well as others), it is customary in high-quality receivers to provide adequate decoupling and filtering of the DC power lines and to provide the LO circuit with a voltage regulator of its own. The voltage regulator must be separate from any others used for the rest of the receiver.

Noise produced by the oscillator itself can be minimized by using a low-noise transistor or IC device, and by proper mounting of the components. A major contributor to phase

noise, for example, is movement of the components. Remarkably small amounts of motion on frequency-determining components, or other components that affect frequency indirectly, can cause FM, PM, or phase noises.

Note that the spectrums shown in Figs. 4 and 5 assume that there are only two frequencies present (F_1 and F_2 , or F_{RF} and F_{LO}). That rarely occurs in real situations. In the radio receiver, for example, there are a lot of signals arriving at the antenna simultaneously. They can cause problems even if they are not situated at the RF and image frequencies (or at their phantoms around the harmonics of the LO). If a group or spectrum of signals appears at the RF input port, then all of them will be converted by the LO. In addition, they will beat against each other and produce new signals that also heterodyne against the local-oscillator signal. This is called intermodulation distortion (IMD).

All in all, with these problems and others, what arrives at the output of the mixer is a real mess for the IF amplifier to handle. Fortunately, there are some things that we can do about it.

Practical Mixer Circuits. The passive Schottky diode double-balanced mixer (DBM) circuit is probably one of the best solutions to the problems normally encountered with mixer circuits, especially if the desired effect is to keep signals not needed out of the IF amplifier. A non-balanced mixer (which is what is in most AM and FM BCB radios) passes the two input frequencies and all of their products to the IF

output (as in Fig. 2). A single-balanced mixer will suppress either F_{LO} or F_{RF} , but not both. A double-balanced mixer, on the other hand, suppresses both F_{LO} and F_{RF} in the IF output, so the output only contains the sum and difference products. A well-designed DBM will also suppress the even harmonics of the LO and RF input signals.

The DBM circuit in Fig. 7 consists of a diode ring (D1-D4) of Schottky devices, and two 1:4 BALUN transformers in which the 4R winding is center-tapped. The LO and RF signals are applied to the ends of the ring, and the IF is taken from the center-tap of T2, the RF input transformer. The degree of suppression of F_{RF} and F_{LO} is determined by the degree of balance. The balance is controlled by the construction of transformers T1 and T2, and by the matching of diodes D1 through D4. Isolation between F_{LO} and F_{RF} is improved by the diode switching action of D1-D4. The switching action acts to prevent transformers T1 and T2 from interacting with each other.

One line of popular commercial DBMs is manufactured by Mini-Circuits (P.O. Box 350166, Brooklyn, NY 11235-0003; Web: www.minicircuits.com). The basic internal circuit is shown at the top of Fig. 8, and the pin-outs are shown at the bottom. Pin 1 is identified by the blue dot around the terminal. There are several models popular with amateur builders, and these are summarized in Table 1. These mixers are designed for a +7 dBm LO signal level and RF signal levels up to +1 dBm.

Figure 9 shows a basic circuit in block-diagram form for using the Mini-Circuits' mixers. In this case the SBL-1 is selected, but the same pinouts are found on others. The RF input signal is applied to Pin 1.

Note the presence of some RF input filtering. It is there to improve the IMD performance. Depending on the application, specifically the RF frequencies involved, the filter might be a low-pass filter, high-pass filter, or bandpass filter. If the range of RF input frequencies is limited, then opt for the bandpass filter so that undesired signals above and below the band of interest are attenuated. Otherwise, determine where the interfering signals are

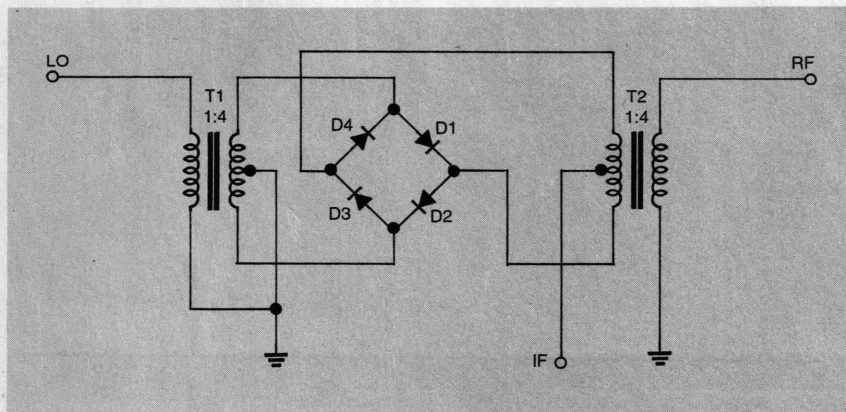


Fig. 7. Here's a generalized schematic for a diode-ring double-balanced modulator.

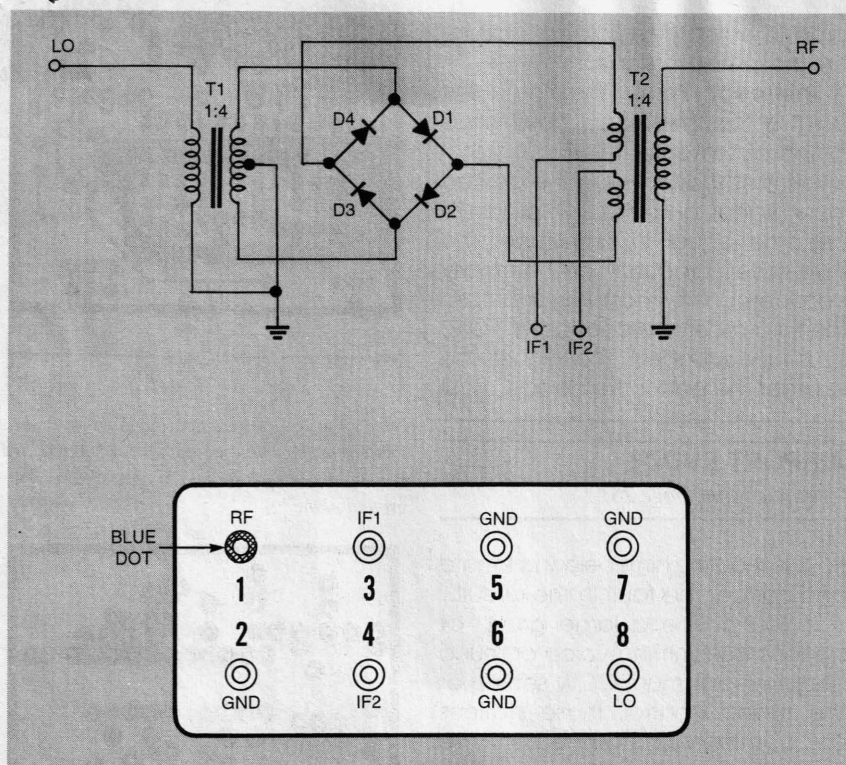


Fig. 8. One popular line of DBMs is made by Mini-Circuits. Here is the basic circuit and pinouts for the members of the series. More details can be found in Table 1.

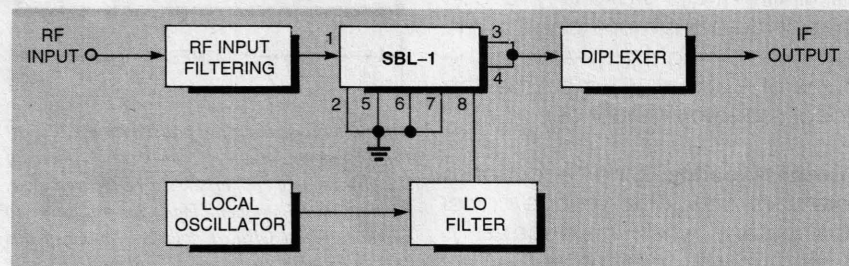


Fig. 9. This block diagram shows how the Mini-Circuits' DBM is typically used.

TABLE 1—MINICIRCUITS MIXERS

Model No.	LO/RF	IF	RF Pin	LO Pin	IF Pin	Ground Pins	Case
SRA-1	0.5-500	DC-500	1	8	3,4*	2,5,6,7	2
SRA-1-1	0.1-500	DC-500	1	8	3,4*	2,5,6,7	2
SRA-2	1-1000	0.5-500	3,4*	8	1	2,5,6,7	2,5,6,7
SBL-1	1-500	DC-500	1	8	3,4*	2,5,6,7	None
SBL-1-1	0.1-400	DC-400	1	8	3,4*	2,5,6,7	None

* Pins must be connected together

with respect to the desired signal and select a type of filter and cut-off frequency that either gets rid of the most signals or the most dominant signal. Whichever filter is selected, however, it must be designed to terminate in a 50-ohm resistive impedance.

If a very pure frequency conversion is necessary (as in a high-quality receiver or converter), then

place a filter in the line between the LO output and the mixer's LO input. As with the RF filter discussed above, select the filter that eliminates the greatest amount of undesired energy. For the most part this means a low-pass filter with a cut-off frequency above the highest LO frequency and the lowest harmonic. Sub-harmonics are also an issue, but are usually less of a

problem. If they are, then either use a bandpass filter centered on the LO range, or cascade a low-pass and high-pass filter.

The local oscillator must be able to produce an output level that will produce a power level of +7 dBm at the LO input (pin 8). The LO output power must be sufficient to overcome the losses of the filters and any other components in the signal path, yet still produce +7 dBm at the LO input of the mixer. For example, if the filter has a 2 dB insertion loss, then the LO must produce +9 dBm output power into a 50-ohm load.

Some designers place a 1- to 3-dB fixed broadband attenuator in each of the signal lines of the mixer. The idea is to damp impedance excursions and to allow the mixer to see a constant 50-ohm source or load impedance at each pin. If you use well-designed filters for the inputs and terminate the IF output properly, then the attenuator should not be necessary. If you prefer to use attenuators, however, the Mini-Circuits' catalog (see their Web site) has examples.

The best way to terminate the IF output of the mixer is to use a diplexer circuit. A diplexer is a circuit that produces a constant impedance over a broad range of frequencies, and will pass a desired frequency and (here is the important part) absorb undesired frequencies. One way to build a diplexer is to connect both high-pass and low-pass filters at the IF output of the mixer. Depending on whether you want the sum or difference frequencies, connect the output of one filter to the load being served (e.g. a following amplifier) and the other to a resistive dummy load that is matched to the system impedance

Diplexer Design. Figure 10 shows a practical bandpass diplexer design intended for use with mixers. The values of the R_O resistors is the system design impedance (50 ohms in most cases). The values of $L1$, $L2$, $C1$, and $C2$ are determined by the center frequency (f_o) of the desired pass band. The Q of the filter is f_o divided by the necessary bandwidth (BW). The values of these components for

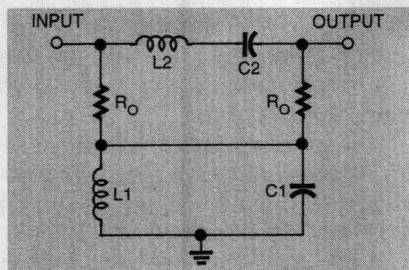


Fig. 10. Here's a practical bandpass diplexer design intended for use with mixers. Values are calculated using the equations in the text.

any given frequency can be calculated from the equations below:

$$Q = \frac{f_o}{BW_{3dB}}$$

and

$$\omega = 2\pi f_o$$

The component values are:

$$L2 = \frac{R_o Q}{\omega}$$

$$L1 = \frac{R_o}{\omega Q}$$

$$C1 = \frac{1}{L1 \omega^2}$$

$$C2 = \frac{1}{L1 \omega^2}$$

Mixer Selection. Several different specifications may be applied to a mixer when making a selection. First you want to select the frequency range. Don't select a mixer that barely covers the frequency you are interested in. For example, if you want to handle an RF input signal of 1.2 MHz, don't select a 1- to 400-MHz device. Select a device with a 0.5- or 0.1-MHz lower end. Keep in mind the RF, LO, and IF frequencies when selecting the mixer for an application. Here are some other specification to keep in mind:

Isolation. The LO-RF, LO-IF and RF-IF isolation tells you something about how much signal will feed through the pathway specified. For example, the LO-RF isolation tells you the amount the LO signal is attenuated when it reaches the RF port. Numbers such as 30 to 60 dB are common.

Dynamic Range. This specification tells you the power range in which the mixer works properly. It is a good idea to obtain the highest dynamic range possible.

-1 dB Compression Point. The input signal level that causes a 1 dB

drop in output level, i.e. a 1 dB increase in conversion loss.

Intercept Point. The intercept point is the theoretical point on a graph of the output-vs.-input signal levels that shows where the desired input signal and the n^{th} products become equal in amplitude. The third-order products are normally considered the most important, so the third-order intercept point (TOIP) is usually specified. The higher the number the better the device. Ω

AIRPORT BUDDY

(continued from page 28)

troubleshooting hints below to locate and correct the fault in the circuit.

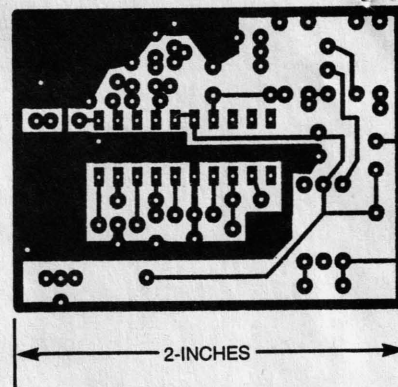
If you notice a large group of stations transmitting voice or music near the maximum CCW setting of the tuning control, those stations are commercial stations that are transmitting at the top of the FM-broadcast band. If you want to eliminate those stations, you can tweak the adjustment range of the receiver by slightly spreading the turns of L1; that will increase the local-oscillator frequency.

Troubleshooting. If the headphones are silent, check the voltage across the battery when the receiver is turned on with a voltmeter. The battery voltage should be at least 8 volts—weak batteries will not drive the Airport Buddy. Also measure the current; the normal value is about 10 milliamperes. The polarized components (D1, IC1, Q1, and the electrolytic capacitors) should be checked for proper orientation. The voltage at pins 5 and 6 of IC1 should be at least 7 volts. If it isn't, D1 might be installed backwards.

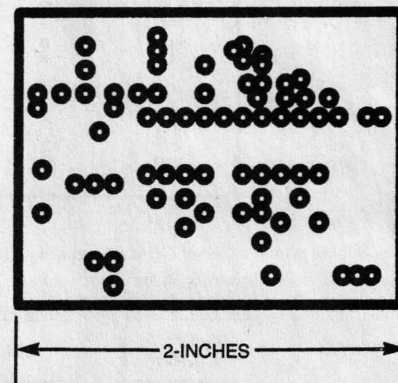
If power to the circuit is normal, check the voltage at pin 7 of IC1. It will normally be about 0.25 volts or less when no signals are being received and increase when an RF carrier is detected.

Check the voltage at the source of Q1. If it is about 1 volt, the transistor is drawing current. If not, check its orientation or try a new transistor.

If it is suspected that IC1 is not operating, carefully check the components that form the local oscillator: R1, C3-C6, L1, and D2. Check the



Here's the foil pattern for the Airport Buddy. The circuit is simple enough for a single-sided layout; the other side is a simple solid-copper ground plane.



Although this is not a foil pattern for the Airport Buddy's ground plane, it shows the locations of the holes where copper should be cleared away so that the component leads do not short to ground. If you wish, a negative image of this layout could be used as a "foil pattern" that will only etch away a ring around the holes that need to be clear.

bias on D2 to be sure that it changes when R6 is turned. Measure the voltage at each pin on IC1 and compare the reading with Table 2. The voltages listed assume that the battery voltage is at least 7 volts. If any pin on IC1 does not seem right, carefully check the wiring for short circuits. If a fault cannot be found, replace IC1.

Using the Airport Buddy. When receiving signals at the airport, adjust the tuning control over its range very slowly to search out any possible transmissions. Once you have found the active frequency of the control tower, you can then keep the tuning control in that general vicinity to hear the audio communications as each aircraft arrives and takes off. Ω

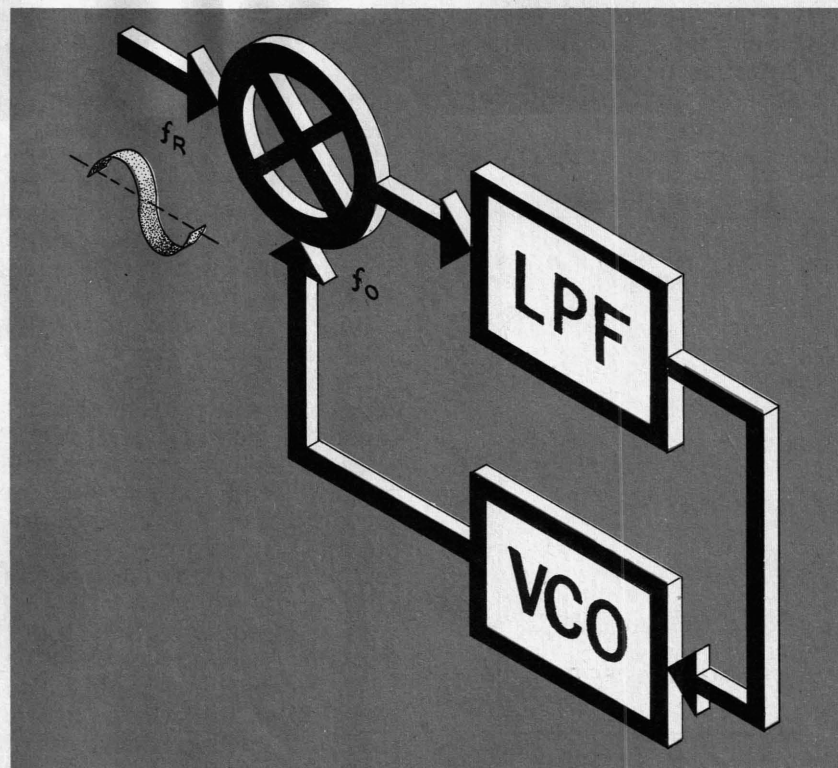
THIS IS THE THIRD ARTICLE IN A SERIES on phase-locked loop and related circuits. It focuses on the monolithic NE565 phase-locked loop and the NE566 function generator, both prime sourced by Philips (Signetics) but alternate sourced by at least four other multinational suppliers. The electrical characteristics of these two PLL ICs are described, and their block diagrams are presented. Circuit schematics based on these devices are included for learning and experimentation.

The last two articles in this series explained the basic operating principles of the phase-locked loop and examined applications for the popular CD4046B PLL IC.

The NE565 PLL IC

The NE565 is a general purpose monolithic PLL used in circuits for frequency-shift keying (FSK), telemetry receivers, tone decoders, and FM discriminators, to mention just a few. Figure 1 shows the principal functional blocks of the NE565: phase detector (comparator), amplifier, stable voltage control oscillator (VCO), and low-pass filter. The self-contained adaptable filter and demodulator cover a frequency range of 0.001 Hz to 500 kHz.

The center frequency of the NE565 is determined by the free-running frequency of the



PHASE-LOCKED LOOPS

Learn more about designing with phase-locked loop circuits in this continuation of last month's article

VCO. This frequency can be adjusted externally with a resistor or capacitor. The low-pass filter, which determines the capture characteristics of the loop, is

formed by an internal resistor and an external capacitor. Figure 2 is the pinout diagram of the most popular plastic, 14-pin DIP package for the NE565.

The NE565 is not as versatile as the CD4046B because the voltage control input pin of its VCO section is permanently tied to the amplifier output through the internal 3.6-kilohm filter resistor (see Fig. 1). As a result, its VCO can't function separately from the rest of the device. While the NE565 works well in many of the same applications as the CD4046B, unlike that PLL device, it is not recommended as a general-purpose signal generator.

Figure 3 is the schematic for an NE565 with external components organized as a signal tracker or FM demodulator. The PLL is powered by a split 12-

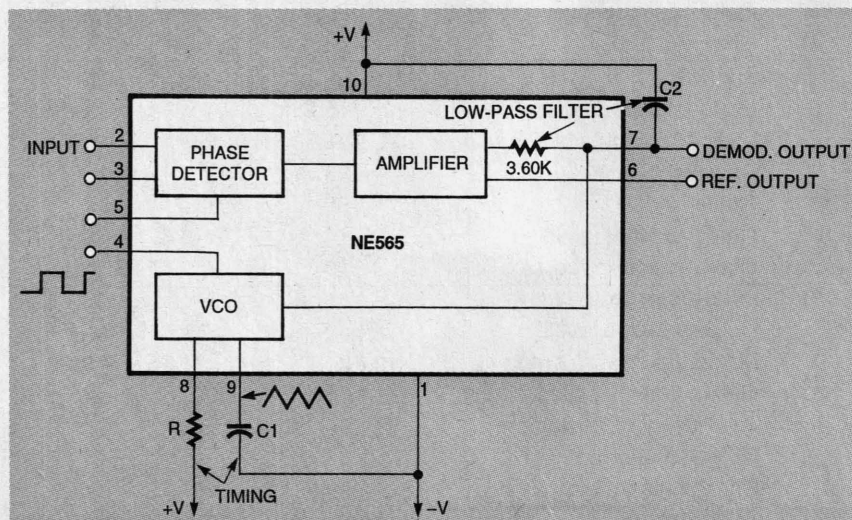


FIG. 1—BLOCK DIAGRAM OF THE NE565 phase-locked loop IC.

volt (+6-volt and -6-volt) power supply. The external signal input to be tracked or demodulated is typically connected to input pin 2 of the phase detector of the PLL, while

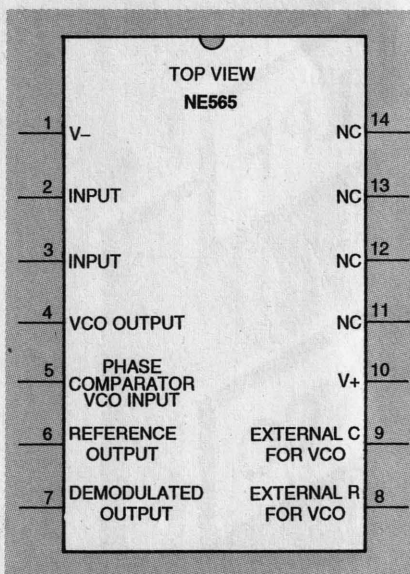


FIG. 2—PINOUT DIAGRAM of the NE565.

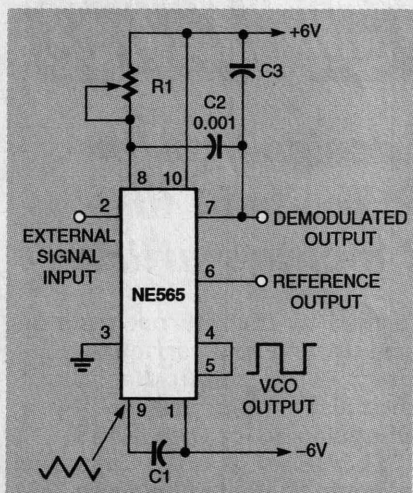


FIG. 3—EXTERIOR COMPONENTS for a signal tracker and FM-demodulator.

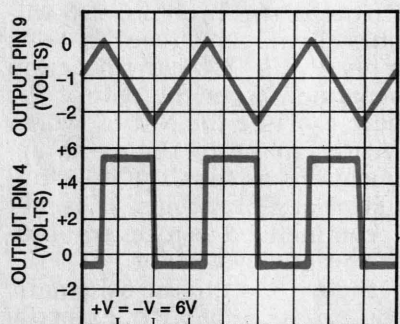


FIG. 4—VCO OUTPUT WAVEFORMS for the NE565 with a plus and minus 6-volt power supply.

unused pin 3 is grounded.

Output pin 4 of the VCO is tied to pin 5, the phase detector's input terminal, to complete the phase-locked loop. The free-running frequency (f_o) of the VCO can be adjusted with the RC network connected to pins 8 and 9 so that it corresponds to the mid-range value of the external input signal.

Under those conditions, the VCO's frequency can "lock" to the input signal. This frequency lock condition occurs because the mean DC level of the phase detector's amplified output is directly proportional to the difference between the input and VCO frequencies, and it can act to voltage control the VCO input.

As a consequence, if the input frequency rises above that of the VCO, the detector's output also rises and automatically drives the VCO's frequency toward that of the input until locking occurs.

The single-pole loop filter formed by capacitor C3 (between pins 7 and 10) and the

internal 3.6-kilohm resistor at pin 7 causes a short time delay. Therefore, if the input signal is noisy, has jitter, or is frequency modulated (FM), the VCO locks to the mean frequency of the input signal and generates a "clean" output at pin 4 or 5. This produces a demodulated FM output at pin 7. The 0.001 μ F capacitor (C2) connected between pins 7 and 8 assures circuit stability.

NE565 characteristics

Table 1 summarizes the leading characteristics of the NE565. The IC is normally configured with a split (positive and negative) power supply, which must be between ± 6 and ± 12 volts, but it can also be powered by a single-ended +12 to +24-volt supply.

The phase detector section of the IC has a typical input impedance of 10 kilohms on each terminal. The IC can lock and track to input impedance of 10 kilohms on each terminal, and lock and track to input signals with amplitudes as low as 1 mil-

TABLE 1—ELECTRICAL CHARACTERISTICS OF THE NE565

Parameter	Min	Typ	Max	Unit
Supply voltage	± 6		± 12	V
Input impedance	5	10		k Ω
Input level for tracking	10			mVRMS
VCO Characteristics				
Center frequency		500		kHz
Drift with temperature		300		ppm/ $^{\circ}$ C
Drift with supply voltage		0.2	1.5	%/V
Triangle wave				
Output voltage level	1.9	2.4	3	Vp-p
Linearity		0.5		%
Square wave				
Logic "1" output voltage	+4.9	+5.2		V
Logic "0" output voltage		-0.2	+0.2	V
Rise time		20		ns
Fall time		50		ns
Output current (sink)	0.6	1		mA
Output current (source)	5	10		mA
Demodulated output				
Output voltage level	4.0	4.5	5.0	V
Maximum voltage swing		2		Vp-p
Output voltage swing		200	300	mVp-p
Total harmonic distortion		0.4	1.5	%
Output impedance		3.6		k Ω
Offset voltage (pins 6 to 7)		50	200	mV
AM rejection			40	dB

livolt RMS. Normally, input signals should be AC coupled, but they can be DC coupled if the DC resistances seen from pins 2 and 3 are equal and there is no DC voltage difference between those pins.

The high stability of the NE565's VCO can be seen in its typical drift with temperature specification of 300 ppm/°C and its typical drift with supply voltage specification is 0.2% per volt when the supply voltage is ± 6 to ± 7 volts. Both of these values are measured at f_o .

The VCO provides well formed, TTL-compatible square waves and triangle waves, as shown in Fig. 4. Those waveforms were obtained with the IC powered by a split 12-volt supply. The square waves, with typical rise and fall times of 20 and 50 nanoseconds, respectively, appear at pin 4, and highly linear triangle waves appear at pin 9 of the VCO.

The VCO's free-running frequency (f_o) is set by trimmer potentiometer R3 in series with the +6-volt power source between pins 8 and 10 and by capacitor C1 in series with the -6-volt power source between pins 9 and 1. That frequency in kilohertz can be calculated as:

$$f_o = 1.2 / (4 RC)$$

where R is in kilohms and C is in micofarads.

Resistor R3 can have any value between 2 and 20 kilohms, but the optimum value is about 5 kilohms. Capacitor C1 can have any value. Normally, the NE565 will phase-lock to any input signal frequency that is within the range of plus or minus 60% of the f_o value. This is known as the circuit's *lock range*.

The output section of the IC yields a demodulated output at pin 7, and pin 6 gives a DC reference voltage that is close to the DC potential of pin 7. If a resistor is placed between pins 6 and 7, the gain of the IC's output stage can be reduced with little change in the DC voltage level of the output. This allows the lock range to be decreased to a value as low as 20% of f_o , with little change in the f_o value.

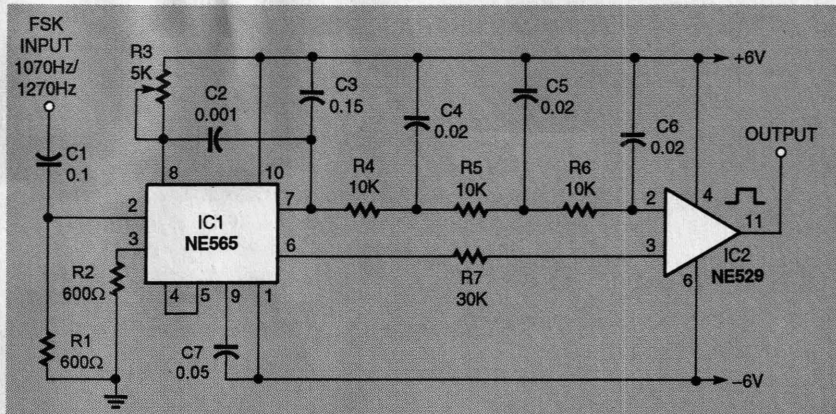


FIG. 5—FREQUENCY SHIFT KEYING (FSK) demodulator circuit.

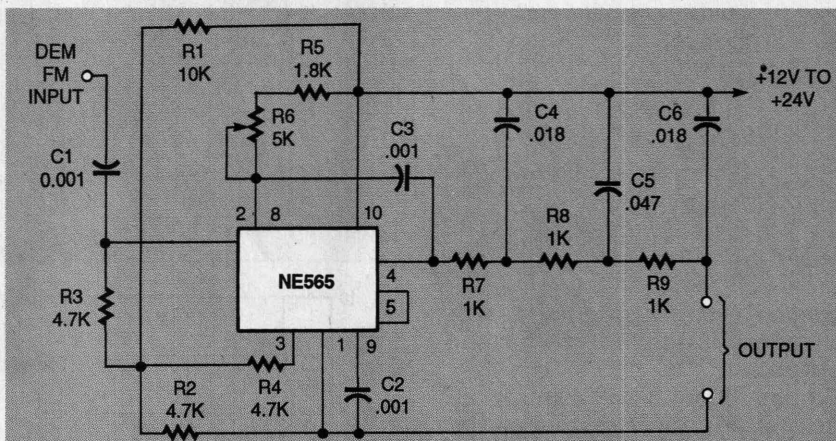


FIG. 6—60 kHz FM DEMODULATOR with a single-ended power supply.

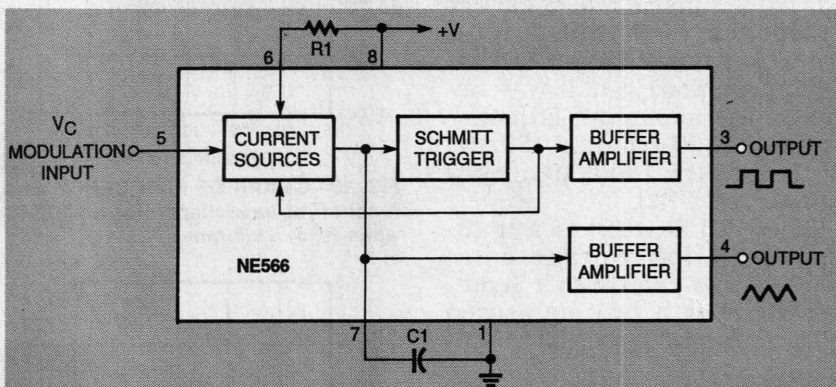


FIG. 7—FUNCTIONAL BLOCK DIAGRAM OF THE NE566 function generator.

FSK demodulator

Frequency shift keying (FSK) is widely accepted in digital communications systems. The transmitter generates a continuous two-tone carrier signal from binary signals with a *mark* or logic 1 state represented by one tone and the *space* or logic 0 state represented by another tone. An FSK decoder in the receiver converts the two-tone carrier back to a binary signal.

Figure 5 shows the NE565 (IC1) organized as an FSK de-

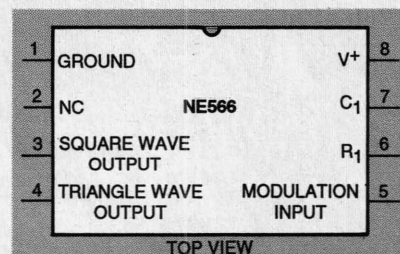


FIG. 8—PINOUT DIAGRAM of the NE566 function generator.

coder of a 1070-Hz to 1270-Hz input waveform. When the signal appears at the input, the loop locks to it and tracks it be-

TABLE 2—ELECTRICAL CHARACTERISTICS OF THE NE566

Parameter	Min	Typ	Max	Unit
Supply voltage	±6		±24	V
VCO Characteristics				
Maximum operating frequency		1		MHz
Drift with temperature		300		ppm/°C
Drift with supply voltage		0.2	2.0	%/V
Control terminal input impedance		1		MΩ
FM distortion(±% deviation)		0.4	1.5	%
Maximum sweep rate		1.0		mHz
Sweep range		10:1		
Triangle wave output				
Impedance		50		Ω
Output voltage level	1.9	2.4		Vp-p
Linearity		0.5		%
Square wave input				
Impedance		50		Ω
Voltage	5.0	5.4		Vp-p
Duty cycle	40	50	60	%
Rise time		20		ns
Fall time		50		ns

tween the two frequencies, with a corresponding DC shift at the output.

Loop filter C2 has a small value to eliminate overshoot on the output pulse, and a three-stage RC ladder filter removes carrier components from the output. This filter has a band edge that is approximately half way between the maximum FSK keying rate (300 baud or 150 Hz) and twice the input frequency (about 2200 Hz).

The NE529 (IC2) is a high-speed analog voltage comparator that combines a Schottky diode with two high-speed TTL gates and a precision am-

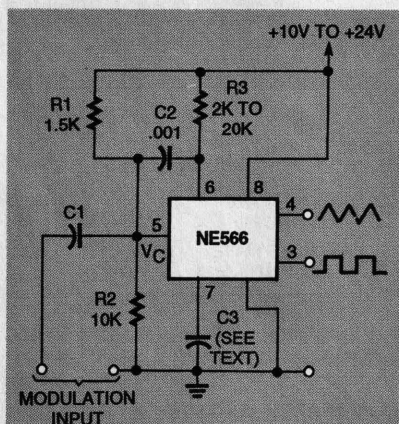


FIG. 9—FIXED FREQUENCY GENERATOR circuit based on the NE566 function generator.

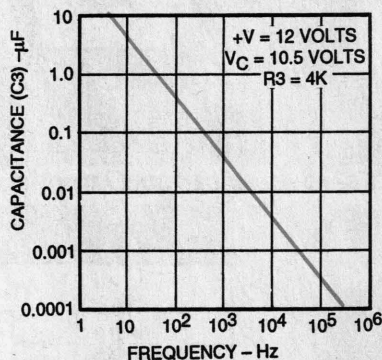


FIG. 10—GRAPH OF FREQUENCY as a function of capacitance for an NE566 when $R_3 = 4$ kilohms.

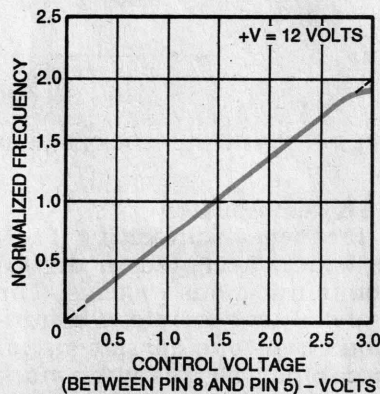


FIG. 11—GRAPH OF NORMALIZED frequency of the Fig. 9 circuit as a function of control voltage.

plifier on a monolithic chip. It is useful in analog to digital conversion and can act as an interface between TTL and ECL

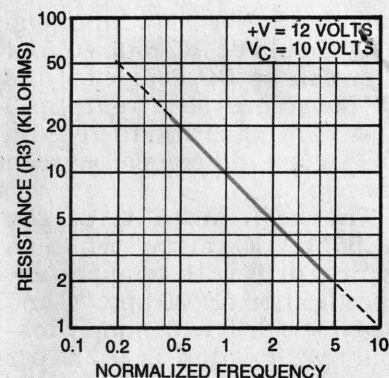


FIG. 12—GRAPH OF NORMALIZED frequency of the Fig. 9 circuit as a function of resistance R_3 .

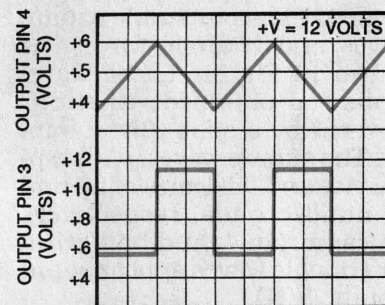


FIG. 13—GRAPH OF VCO OUTPUT waveforms for the generator circuit of Fig. 9.

circuitry.

An NE529 between pin 6 of IC1, the NE565, and the output of the circuit makes the output of the filter CMOS logic compatible. Adjusting trimmer potentiometer R_3 sets the free-running frequency of the VCO to give a slightly positive output voltage when a 1070-Hz input signal is applied.

The input connection of the circuit shown in Fig. 5 is typical for applications where a DC voltage is present at the source, preventing a direct connection. Both input terminals are returned to ground with identical resistors. In this circuit, the values of resistors R_1 and R_2 were selected to give a 600-ohm input impedance.

Single-ended power

Figure 6 shows the NE565 configured as a 60-kHz FM demodulator that is powered from a single-ended 12- to 24-volt supply.

A resistive voltage divider (R_1 and R_2) and R_3 and R_4 apply a balanced bias voltage to input pins 2 and 3 of the NE565. The

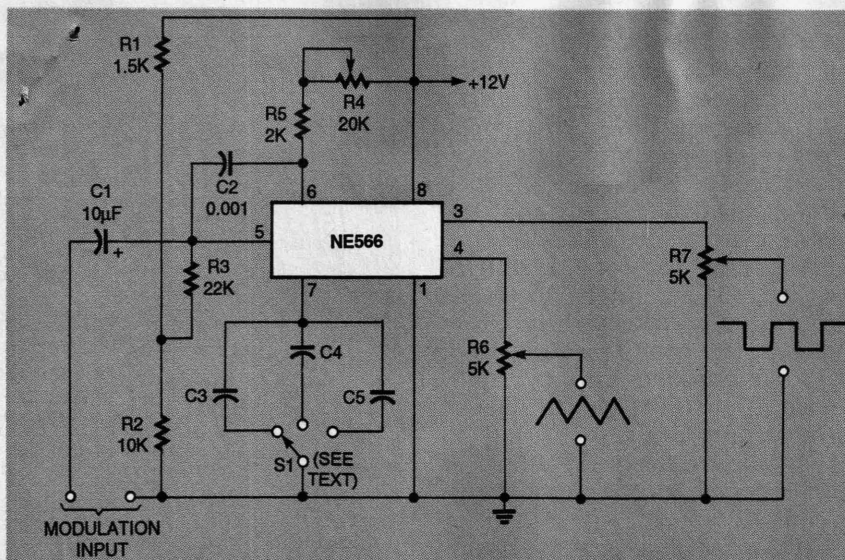


FIG. 14—THREE-BAND FM GENERATOR based on the NE566 general-purpose function generator.

60-kHz FM input signal is AC-coupled to pin 2. The VCO's free-running frequency is set to 60 kHz with R6, C2 and trimmer potentiometer R5. The decoded output signals are fed through a three-stage, low-pass filter to minimize the effects of unwanted noise on the signals.

NE566 fundamentals

The NE566, a general-purpose function generator, is a linear voltage-controlled oscillator capable of producing buffered squarewave and triangle-wave outputs, at fixed and variable frequencies, up to a maximum of about 1 MHz. Its frequency of oscillation is determined by an external resistor and capacitor and the voltage applied to the control pin.

The oscillator can be programmed over a ten to one frequency range by the appropriate selection of an external resistance. It can also be modulated over a ten to one range by a control voltage. The NE566 finds applications in tone, signal, and clock generators, as well as FM modulators.

Figure 7 is a block diagram of the NE566. The principal blocks are current sources, a Schmitt trigger, and two buffer amplifiers. Two essential external components, R1 and C1, are shown. The VCO section consists of the pair of voltage-controlled current sources that linearly charge or discharge an

external timing capacitor.

A linear triangle wave is generated across the capacitor and appears at pin 3, and a square wave generated at the Schmitt trigger switches the current sources when the capacitor voltage reaches preset levels. These waveforms are available at the output pins of the buffer amplifiers.

Figure 8 is the pinout diagram for the popular 8-pin plastic DIP package. The NE566 is, however, also packaged in a 14-pin DIP package. Table 2 lists selected electrical characteristics of the NE566. It can be seen that the impedance values for both the triangle-wave and squarewave outputs are a low 50 ohms.

Figure 9, a schematic showing the NE566 configured as a fixed-frequency FM waveform generator, shows the locations of the external components required for its functioning. Frequency is determined by resistor R3 with capacitor C3 and by the voltage applied to its control terminal.

Resistor R3 must have a value between 2 and 20 kilohms but capacitor C3 can have any value. The control voltage must be between $\frac{3}{4}$ and the full supply voltage.

The output frequency can varied or modulated over a ten to one range by variation of the control voltage. Capacitor C2, connected between pins 5 and

6, has a value of 0.001µF to stabilize the circuit.

The operating frequency of the NE566 is:

$$f_o = \approx (+V - V_C)/R3 \times C3 \times +V$$

Resistor R3 should be in the range of 2 to 20 kilohms.

Figure 10 is a graph of output frequency as a function of the values of capacitor C3 when R3 has a value of 4 kilohms. The frequency can be varied from 5 Hz with a C3 of 10µF to about 200 kHz with a C3 of 0.0001µF. Figure 11 is a graph of normalized frequency as a function of control voltage, and Fig. 12 is a graph of normalized frequency as a function of resistance R3. Figure 13 is a graph of VCO outputs of the circuit in Fig. 9 when the power supply is 12 volts.

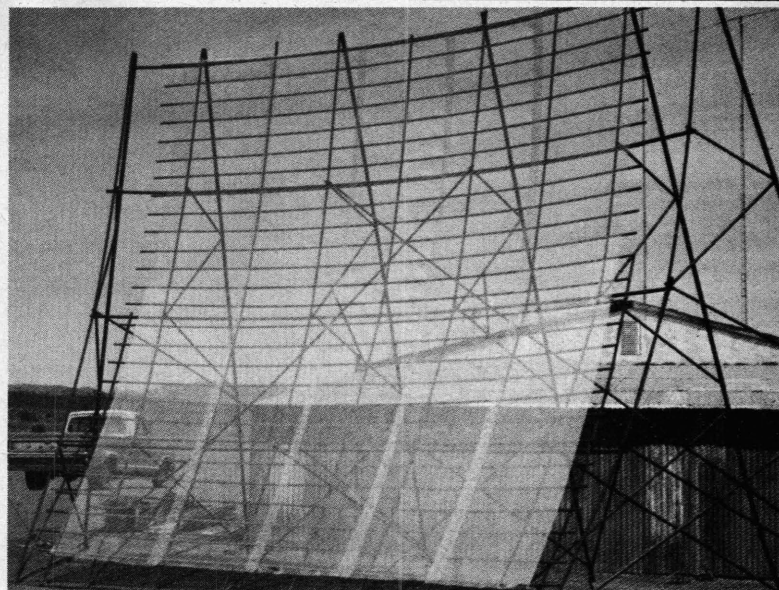
Figure 14 shows a modification that can be made to the Fig. 9 schematic to convert it into a wide-range, three-band FM generator. The frequency can be altered by trimmer potentiometer R4, and three different frequencies can be obtained with selector switch S1 in series with C3, C4, and C5 in parallel and connected to pin 7 of the NE566. As stated earlier about Fig. 9, capacitor C3 can have any value, so C3, C4, and C5 can have values to give three different frequencies.

The voltage levels of the triangle-wave and squarewave outputs can be varied by trimmer potentiometers R6 and R7. Those waveforms can be frequency modulated by applying the modulation waveform to pin 5 through input capacitor C1. Resistor R3 raises the circuit's input impedance to its value of 22 kilohms.

The NE565 PLL and associated NE566 function generator are useful integrated circuits to be familiar with. The circuits presented here should give you a working knowledge of the IC's applications.

The final article in this series about PLL devices will describe the NE567, a highly stable phase-locked loop device. Although it is primarily used as a tone decoder, the IC has additional applications as well. □

Home Reception Using Backyard Satellite TV Receivers



SWAN SPHERICAL ANTENNA is effective in receiving satellite transmissions and can be built relatively inexpensively.

Part 4—In this installment of a series, we will go into more technical details on receiver characteristics and specifications and will show how some satellite receivers have been built at comparatively low cost.

ROBERT B. COOPER, JR.

IN PARTS ONE, TWO, AND THREE OF THIS multiple-part series (appearing in the August, September, and October 1979, issues of **Radio-Electronics**) we learned how the geo-stationary satellite system is designed, what it is intended to do and what a private individual, living someplace south of the 80th north parallel, north of Venezuela, and east-west between Bermuda and Hawaii can anticipate being able to receive with a private, backyard satellite television terminal. Satellite television is the next "generation" of television service in America and throughout wide areas of the world. Because of the mechanics of the service, it is virtually immune to interference and signal degradation, is not adversely affected by weather, and holds the potential to provide every home in North America with several hundred direct-access television channels!

Receiving system

Having determined that the basic system consists of an antenna, a low-noise amplifier (LNA) and a receiver-demodulator, let's look at what it is that goes into each of these three major component modules to make up the operating system.

The antenna system has been adequately covered in previous portions of this series. Basically, in order to achieve

the kind of gain necessary (38 to 45 dBi) a parabolic reflector is the best antenna choice. This parabolic reflector has a single focal point where all of the energy intercepted by the reflective antenna surface is re-directed and focused. There are several acceptable members of the antenna family known as parabolics that can be pressed into this service; prime focus parabolics, Cassegrain parabolics and spherical parabolics are included. For as long as the (limited) supply holds out, surplus (as in no longer used in commercial or military service) parabolic (or "dish") antennas larger than 8 feet in diameter provide very economical "reflector surfaces" for most portions of North America. The exception to this is in New England where anything smaller than a twelve-foot reflector surface would be a mistake. Beyond that, one of the least expensive antenna surfaces for this service has been developed by a fellow in Arizona named Oliver Swan. Using aluminum window screening as a reflector surface, and stock square aluminum or steel tubing as reflector frame material, Swan has developed a spherical antenna system that can be constructed in virtually any size from 10 feet by 10 feet to 20 feet by 20 feet for as low as approximately \$500 for the ten-foot by ten-foot version. It is inevitable that some commercial firm will soon begin marketing antenna "kits" in this

area, perhaps copying the Swan developed spherical antenna and that from this will spring a whole new family of "backyard decorative pieces."

Because we are dealing with a low-power transmitter source (the typical satellite has a 5-watt peak power transmitter) and a fairly high loss between the "bird" and your receiving location (196 to 200 dB is typical at 4 GHz), not very much signal power arrives at your antenna. Fortunately, the signal received is very constant (variations of ± 0.7 dB over a full year are typical limits) and this allows us to design the system for peak performance and forget it rather than be concerned with wide-range AGC systems to cope with large signal fluctuations.

To make the most of the weak signal, we have to place a very high gain, and extremely low noise (figure) signal amplifier (or booster in TV terms) right at the antenna. Since the reflector surface on the parabolic is merely a focusing tool, the actual "pickup antenna" is really separate and distinct from the reflector. This receiving antenna, directed backwards away from the satellite and towards the focused energy coming from the reflector surface, is called a "focal point" or feed-point antenna.

The most efficient feed antenna is one that looks at the reflector surface in such a way that the "pattern" on the feed antenna is down 10 dB at the out-

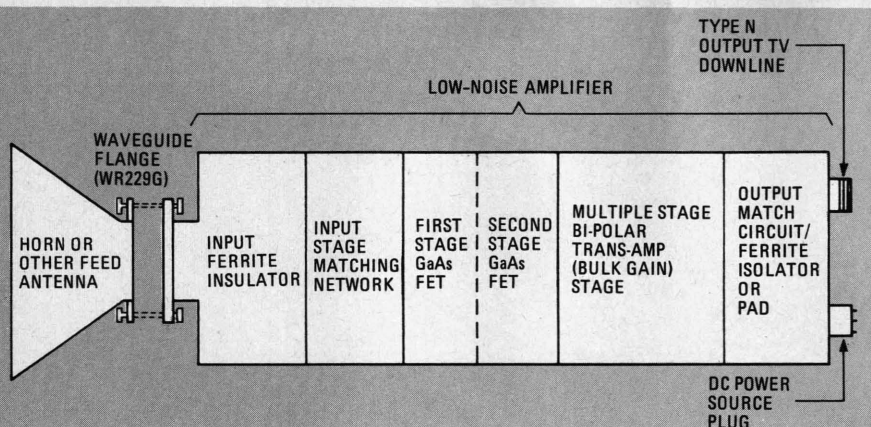


FIG. 1—MOST COMMERCIAL FEED-ANTENNA LOW NOISE AMPLIFIERS consist of a feed antenna (often a horn as shown for a prime-focus feed) which is equipped with a waveguide flange that bolts it directly to the input flange on the LNA (Low Noise Amplifier). The LNA has an input (ferrite) isolator both for frequency selectivity and as an impedance matching device for the first stage of the amplifier itself. The extremely low-noise amplifiers develop their superior operating characteristics because of extreme care taken in matching the first stage to the input source impedance and by carefully hand selecting the expensive GaAs-FET (Gallium Arsenide Field Effect) transistors. GaAs-FET devices are chosen for the first two stages. Once the noise figure is "established" by these two stages, less costly bipolar transistors are used in 3-4 additional stages for "bulk" gain. The output, to the low-loss downline coaxial cable, is through another ferrite isolator device or through a "loss pad" inserted to force an impedance match.

side edges of the reflector's surface area. A horn-feed antenna, properly designed, handles this function. Look closely at Fig. 1. Note that the horn-feed antenna is flanged or bolted directly to the low-noise amplifier itself; the energy from the horn feed-point antenna couples through the waveguide flange into the input circuit on the low-noise amplifier, a section that has a piece of ferrite (rod) in it as an isolator.

Low-noise amplifiers

This commercial style low-noise amplifier is the state-of-the-art high-dollar approach to the low-noise amplification aspect of the system. There are less expensive ways to go as we shall see in subsequent portions of this series. The purpose of the ferrite isolator is primarily to insure that the input circuit to the first active (transistor) amplifier stage sees a constant impedance or load. This is done to insure that the transistor used in the first stage, a GaAs-FET (for Gallium Arsenide Field-Effect Transistor), is noise-figure matched at the 4 GHz operating frequency. Most of the high-dollar GaAs-FET's available for this service have two separate peak operating points; maximum gain does not coincide with best (i.e. lowest) noise-figure performance. In this case, gain is backed off in the first couple of stages as a trade off for lowest noise figure since noise generated in the early pre-amplifier stages is impossible to eliminate later on in the system.

Most of the commercial LNA units employ a pair of ultra-low-noise GaAs-FET's in the first two stages, and then follow that up with between three and five less expensive (typically bipolar as opposed to GaAs-FET) amplifier stages. Once the noise figure for the LNA is es-

tablished by the first couple of stages, less expensive (and higher noise figure) bipolar stages can make up the remainder of the LNA system gain required.

Noise figure is measured in both dB and by the Kelvin noise temperature scale. Most of the commercial data sheets will specify Kelvin temperature only and most commercial installations are using amplifiers with 120-degree Kelvin (or 1.5 dB noise figure) specs.

State-of-the-art has been catching on quickly in this field; in late 1976 the price for a 120-degree Kelvin LNA was in the \$3,500 region. By late 1978 you could find the same amplifier for around \$1,800. Today the price is down in the \$1,000 region and many expect it to drop down close to \$500 by this time next year. That still may be high for your pocketbook and there are other options.

As previously discussed in this series, you can get a raw signal input to the receiver by one of two techniques; use a big antenna and an LNA with not such hot specs, or, use a smaller antenna and a hot-spec LNA. If you set out to build your own antenna system, rather than buying commercially, you might be better off in this fast-changing technology time to invest in a little more steel and mesh and build a larger antenna going in, especially if you plan on having to purchase your LNA.

As recently as early in the past summer anyone who wanted to build his own LNA was pretty much stuck with working with 300-degree Kelvin type bipolar transistors. The belief was that any home constructor attempting to work with the touchy, and hard-to-make-work GaAs-FET amplifiers was probably asking for a quick way to lose a \$100 bill; that being the going price for the GaAs-FET transistors these days

from Hewlett-Packard (HFET 2201). However, during the recently completed Satellite Private Terminal Seminar this past August 14-16 in Oklahoma City, several home terminal builders demonstrated two- and three-stage GaAs-FET amplifiers they had constructed for between \$225 and \$350 that gave an excellent account of themselves against commercial amplifiers costing three to four times as much. This major achievement has changed the name of the private or backyard terminal game for the home constructor.

Well now; if a chap in Arizona can build a 10-foot by 10-foot spherical reflector and feed for \$500 or less, and you can build your own GaAs-FET low-noise amplifier on the kitchen table for \$350 or less, that starts to get the home-constructed terminal down to an affordable price does it not? What about the receiver?

Receivers

In 1976 the first satellite video receivers around came into the cable television field via the Intelsat or international satellite marketplace. They cost upwards of \$10,000 and were literally hand wired and hand aligned.

By early 1978 the price for essentially the same receiver was down to about half that; perhaps \$5,000. But there had been only minor changes in the original design. The price reductions were largely due to slightly more volume production, and of course competition.

Needless to say many people were working on bringing the cost down; way down. Most however were involved in the cable TV, broadcast TV and other commercial market areas where nobody really expected receiver prices to drop much below say \$3,000 for many years to come. Outside of these broadcast related industries other engineers with a totally different set of markets in mind were quietly doing their own developmental work. Their goal was a \$3,000 complete terminal; including the antenna and the LNA.

By mid-1979 some inter-receiver marrying had taken place. Commercial receivers are available in two formats; some tune only one channel and to change channels you have to either change crystals or go through some sequence of screwdriver adjustments, or both. Not exactly what the home viewer accustomed to detent tuning has in mind. The other commercial receiver format is called "frequency agile" and that means you push buttons or twirl a knob and the full set of 12 (or 24) satellite channels flips by in front of you. By mid-1979 some of the commercial receivers in the single channel format were down under \$2500 list price while the tuneable versions were just a tad above \$3,000.

Let's stop for a minute and study Fig. 2. To appreciate what is involved in a satellite television receiver, we ought to understand what it has to do.

In a commercial installation the LNA (which mounts at the antenna, usually married to the feed-horn or focal-point antenna) has to develop sufficient RF signal voltage gain, at 4 GHz, to (1) drive the microwave signal through the interconnecting coaxial cable and into the receiver, and, (2) provide sufficient signal gain to establish the noise figure of the LNA as the noise figure for the whole receiving system.

The typical satellite TV receiver has a relatively high noise figure; 10-12 dB is not uncommon. To attempt to use such a high "front-end" noise figure to receive the weak satellite signals would be a mistake. To lower the noise figure to a more usable level (such as under 2 dB) requires not only a low-noise LNA but sufficient gain in the LNA stages to override the noise contribution by the 10-12 dB noise figure of the receiver. As a rough rule of thumb you need between 2.5 and 3 times as much voltage gain (in dB) as the noise figure (also in dB) to establish the new, lower noise figure of the LNA as the noise figure of the system as a whole.

Back now to Fig. 2. To keep unwanted energy out of the receiver (and there is plenty of unwanted or off-frequency energy floating around microwaves these days) the typical commercial receiver has a pre-selector (either totally passive or active plus passive) at the input. This is followed by a "high frequency-mixer" that combines the incoming (3.7 to 4.2 GHz) signals with a local oscillator signal source generated within the receiver to produce a new lower frequency (IF) output. Gain is

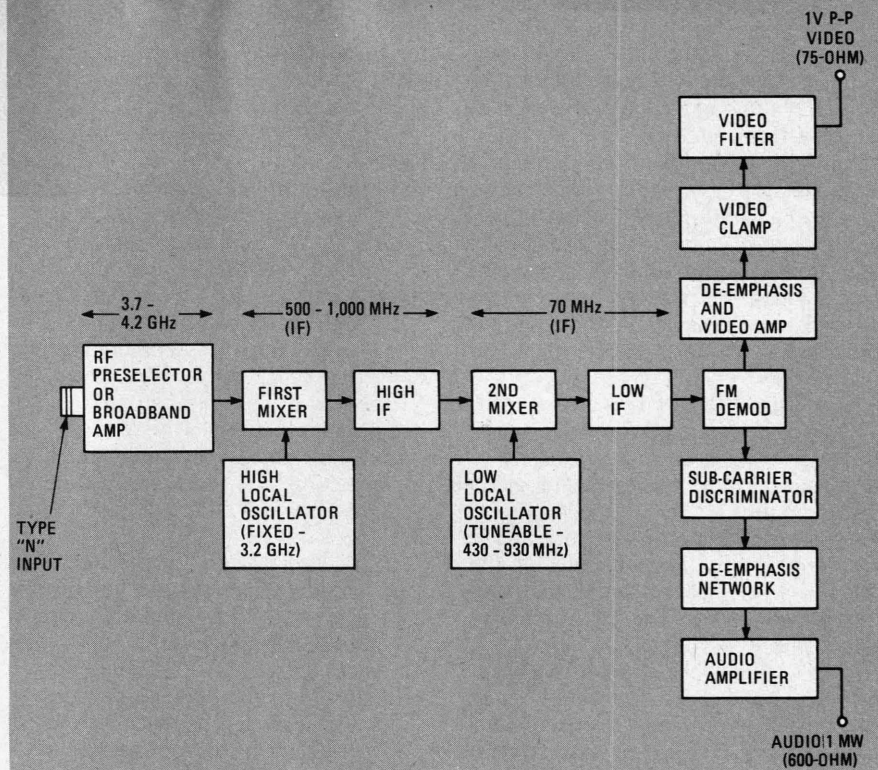


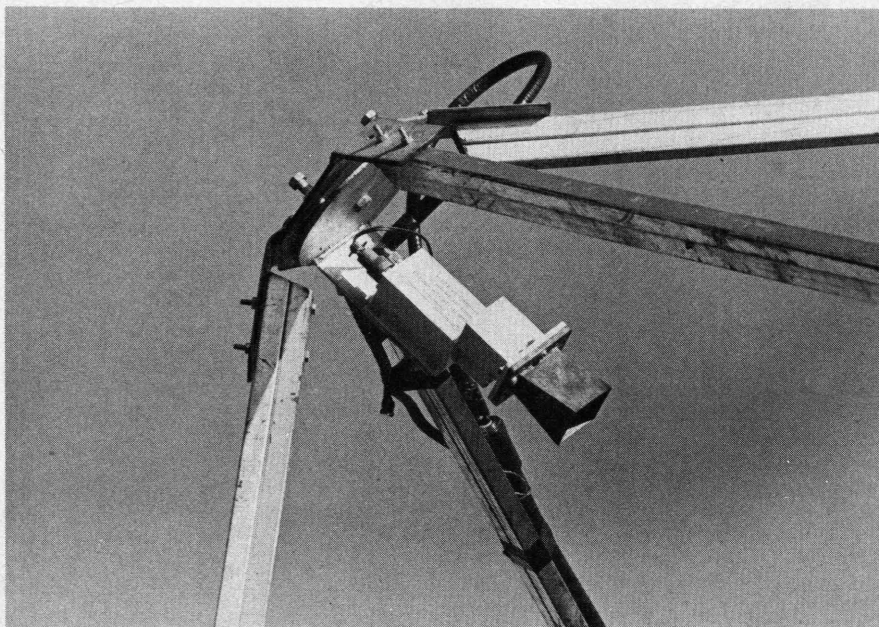
FIG. 2—MOST COMMERCIAL RECEIVERS use high-gain (50 dB) LNA's mounted at the antenna capable of driving 4-GHz region signals through several hundred feet of 7/8ths inch line to the indoor receiver. Typical receiver has either double conversion (approach shown here) or single conversion directly to IF. In receiver design shown, a 3.7- to 4.2-GHz preselector often has enough gain to bring the input RF level up to an adequate voltage to drive the first mixer. Relatively high-level high-frequency local oscillator (3.2 GHz is shown in example) may measure +10 dBm or more. With local oscillator on low side, high IF of 500-1,000 MHz is attained to drive second mixer that down converts to 70-MHz low IF. Oscillator for second local oscillator is voltage- or capacitive-tuned to produce 70-MHz low IF with inputs over full 500-1,000 MHz range. FM demodulator (discriminator) produces basic baseband signal that processes up for video to a de-emphasis circuit and video amp, then to a video clamp to eliminate the 30-Hz energy dispersal waveform, and then to a filter circuit. Audio processes down through separate (6.8 MHz) discriminator, de-emphasis network and audio amplifier.

then applied at the high-frequency IF and then the signal goes through yet a second mixer that further down-converts the high IF to yet a lower IF. This lower IF is often 70 MHz although there are some variations to this rule in

commercial receivers. When we finally reach the lower IF, we have gone through a pair of down conversions each employing a high-quality mixer and a high-quality local oscillator. If this is a frequency-agile (i.e. tuneable) receiver the first mixer is driven by a tuneable local oscillator source while the second mixer is driven by a fixed local oscillator source. Just for dollar reference, we are looking at using \$75 to \$100 mixers in these applications and the local oscillators are priced in about the same range. If this suggests that microwave components or modules are not cheap, you read the message correctly.

Once at the low IF we are ready to go to work on the modulation itself. Gain at a relatively low IF such as 70 MHz is inexpensive these days and 40-50 dB of gain in this range is typical. When the twice-down-converted signal is built up to a sufficient voltage level, it is ready to be demodulated. Remember that the video is frequency modulated onto the carrier, and the audio coming along with the video is further frequency modulated as a sub-carrier. This says that we use discriminators to demodulate the video and the audio in our detection system.

By removing the video signal out of the IF signal with a detector, we end up



HORN ANTENNA/LNA combination points directly towards the dish antenna. Coaxial cable is used to connect the LNA to the receiver.

with what is called baseband; that means pure video in this case. Only because the audio is carried along as a 6.2- or 6.8-MHz add-on or subcarrier, when we demodulate to baseband video we also have a aural subcarrier in the baseband output. By using a low-pass filter for the video and a high-pass filter for the aural subcarrier, we can then separate the video into one chain for further processing and the audio into another.

The video is preemphasized at the uplink transmitter site as a means of increasing the system performance and at the receiver we need to deemphasize to establish the original baseband video characteristics. The deemphasis network is strictly an L-C network and is not complicated. Next in line for the video is a video clamp circuit that may mystify you if you are accustomed to normal video techniques.

Video at the uplink site is "frequency-dithered" or dispersed at a 30-Hz rate as a means of reducing potential interference between strictly terrestrial 4 GHz video circuits (such as the telephone companies employ) and the satellite service. The easiest way to clean up

the dispersal waveform is to shove the video through a clamping circuit. If you clamp something like this *hard* enough, the 30-Hz waveform simply goes away. Finally a bit of passive video filtering and you are in business with baseband video (typically 1 volt peak-to-peak).

Over on the audio side, after passing the 6.2 or 6.8 MHz aural subcarrier through a frequency filter that eliminates the video baseband information, the signal is fed into yet another discriminator (detector) that recovers the audio. From here it goes through yet another deemphasis network (this one for the audio) and finally an audio amplifier. Most commercial receivers release the audio across a 600-ohm balanced output line.

If you are engaged in the television receiver servicing industry, you may be asking yourself why this should cost between \$2500 and \$3500 a pop. If you are new to receivers in general, you probably have the opposite reaction.

As we shall see in the next part of this series of articles, several experimental or private terminal builders have asked themselves the same thing. One terminal builder, Taylor Howard

of California, has managed to assemble the LNA (a bipolar unit in his case) and the receiver for around \$1,000. He did this back in 1976-77 when parts were considerably more expensive and we estimate you can do it today for under \$700.

Assuming you don't want (or need) to start off with a bag full of new parts, and can assemble some equipment from other services into a satellite TV receiver, just how simple can it really be? Well, a man in South Carolina by the name of Robert Coleman has put together a 10-foot dish, a two-stage GaAs-GET LNA and a complete receiver for around \$500! His "secret," if you can call it that, is that he is a sharp attendee of Hamfests and other outlets where *surplus* electronic equipment is brought out for sale at often just a few pennies on the original dollar value. The Coleman approach is a good one, but it requires being able to trace down surplus parts, modules and components that may not be a good supply because of limited production runs many years (or decades) ago. Still, if this approach does interest you and you are not afraid to go into the surplus market to look for parts, there is help available for you in this specialized area.

Suppose you wanted to try a cross between building a complete terminal receiver from scratch and assembling one from surplus equipment? Well, that is an approach many people have followed, largely patterned after the work done by English satellite TV experimenter/pioneer Steve Birkill (amateur G8AKQ). The Birkill receiver is similar to that shown in Fig. 3. The LNA is a bipolar system of three to five stages using Hewlett-Packard HXTR (6102 and 6101) transistors. For those who want to investigate this particular approach, Hewlett-Packard Application Note No. 967 tells how to build a stage of this amplifier at 4 GHz (a multiple stage-device is simply several separate stages cascaded together).

The Birkill Receiver places the LNA stages at the feed antenna, follows that with a double-balanced mixer (also located at or near the feed) and the mixer is driven by both the input 4 GHz range signal(s) plus a "free-running" oscillator operating at around 3,200 MHz. There are several ways to derive the local oscillator injection signal; one of the easiest is to use a completely self-contained oscillator. One of the 8360-family of oscillators manufactured by AvanteK, Inc., 3175 Bowers Ave., Santa Clara, CA 95051 will do the job nicely. This TO-8 packaged device has four pins on it; one for the positive operating voltage, another for a ground, a third for the RF output in the gigahertz region and a fourth for a tuning voltage that allows you to run the oscillator through a 500-MHz span. Most homebrew (from

LEARNING MORE ABOUT SATELLITE TV

Readers with an interest in pursuing further the details of the Home Satellite TV Reception system may be interested in the following reference materials available:

Satellite Study Package—includes a 72-page "Handbook" describing the complete home satellite TV receiving system, how the satellites operate, where to locate equipment and what services are available via satellite, written by Bob Cooper. Also includes 22-by-35 inch four-color two-sided wall chart depicting each of the geostationary TV satellites in operation and explaining the technical equations of the satellite TV system. Satellite Study Package is excellent "starter set" towards a full understanding of the system and getting your own system underway. Price is \$15 via first class mail (U.S. and Canada in U.S. funds) or \$20 elsewhere from Satellite Television Technology, P.O. Box 2476, Napa, CA 94558.

Home Satellite Construction Manuals—three separate manuals available describing construction details (parts lists, sources for parts, schematics where applicable) of two separate receiver systems, low-cost low noise amplifiers and the unique new *Swan Spherical TVRO* antenna. The *Coleman TD-2 Manual* describes conversion of surplus microwave equipment to TVRO terminal plus showing latest low-cost methods of amplifying and converting 4-GHz signals to 70-MHz IF; the original Coleman system cost under \$500 to build. The *Howard*

Terminal Manual describes a 24-channel frequency-agile (tuneable) receiver plus bipolar LNA using off-the-shelf new parts that can be duplicated for under \$800. Taylor Howard, a professor at Stanford, has had this system operational since 1976. The *Swan Spherical TVRO Antenna Manual* describes construction of a unique multi-satellite reflector surface plus a unique feed-horn; inventor Oliver Swan shows how a 10- to 12 foot-size antenna can be assembled with common hardware for under \$300. Price of each manual is \$30 or all three for \$80 from: Satellite Television Technology, P.O. Box G, Arcadia, OK 73007.

Satellite Private Terminal Seminar—this past August more than 500 satellite TV enthusiasts attended first-ever "international seminar" as more than a dozen instructors (including Taylor Howard, Robert Coleman, Oliver Swan, and Paul Shuch) taught in classroom sessions the latest in low-cost satellite TV circuit technology. SPTS '80 will be held in Miami, Florida on February 5, 6 and 7 with special emphasis on how to market and install private low-cost terminals throughout North, Central and South America. There will be exhibits, live-satellite systems demonstrations, and course study books. Registration is \$150 per person; advance registration required. For full information write: SPTS '80/Miami, P.O. Box G, Arcadia, OK 73007 or call (405) 396-2574.

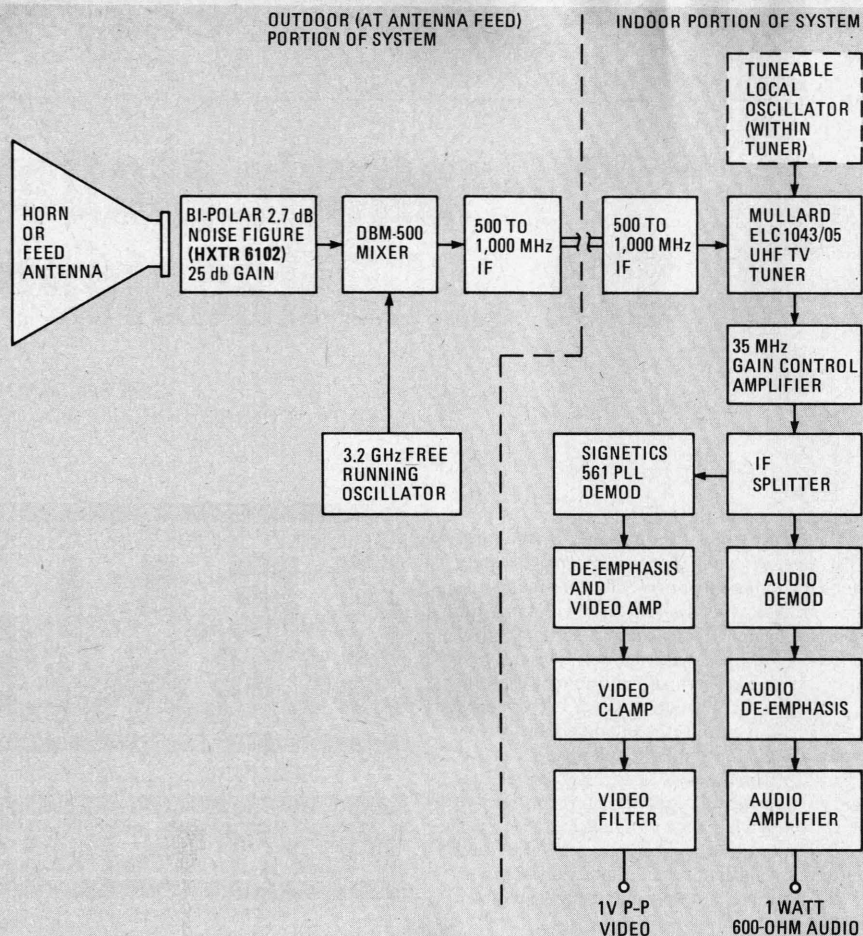


FIG. 3—MOST HOMEBREW TERMINALS place a lower grade LNA stage plus first mixer, local oscillator source and an IF stage or two at the feed antenna, coming down to the baseband demodulator through lower-cost 50-ohm cable at the high-IF (500-1,000 MHz) region. In this version, essentially patterned after English experimenter Steve Birkill, a Mullard ELC1043/05 (European) TV tuner is slightly modified as combination oscillator and mixer to translate high IF range down to 35-MHz region low IF. Birkill processes his 35-MHz IF to video through a Signetics 561 phase-locked-loop demodulator; a system that offers advantages for weak input level signals. Full block diagram is not shown at this time.

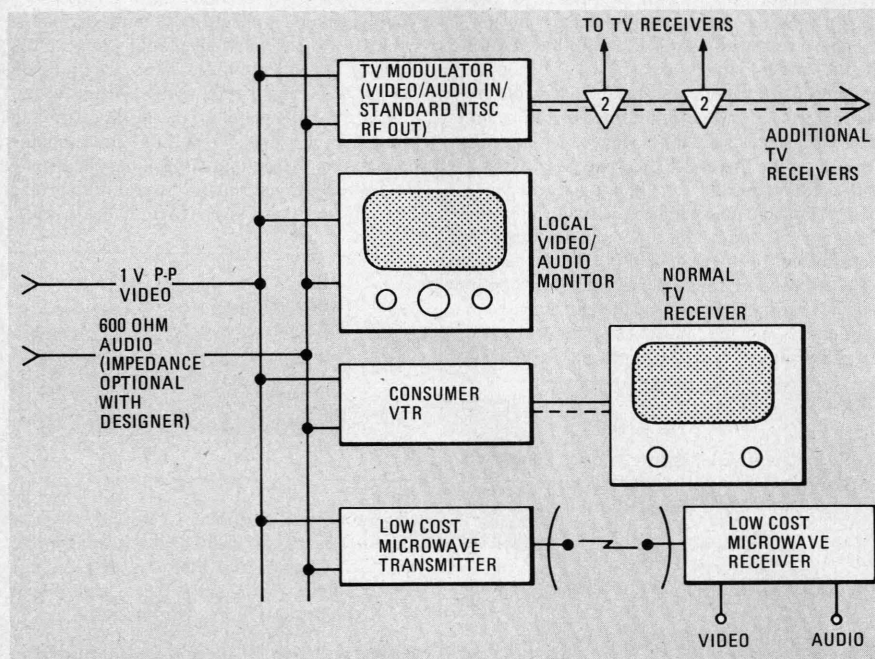


FIG. 4—VARIOUS METHODS OF DISPLAYING BASEBAND (i.e. demodulated) video and audio. Typical receiver produces 1V-P-P video and some usable level of audio (often at 600-ohms balanced although that is design decision of builder). Baseband signals will directly drive TV channel RF modulator, a high-quality video monitor (with audio display system built-in or separate), a consumer VTR (for recording or as loop-thru to use RF modulator), or low-cost (private) microwave system.

scratch) satellite TV receiver builders are using this approach because it eliminates all work at microwave frequencies in deriving the high local oscillator signal.

The Birkill Receiver approach, modified slightly, is shown in Fig. 3. As you can see, the LNA, the high-frequency local oscillator and mixer, plus a bulk gain stage operating in the 500-to-1,000 MHz region is mounted outside at the antenna. This simply means that what you feed "downstairs" to the remainder of the receiver, through coaxial cable, is (relatively speaking) low-frequency signals; in the 500-to-1,000 MHz region in this case. If the run is 100 feet or less, you can get by at these frequencies with RG-8 type coaxial cables whereas a similar run at 4 GHz requires 7/8-inch air-dielectric cable and special fittings.

Once indoors, the Birkill Receiver approach treats the signals contained in the 500-to-1,000 MHz IF bandwidth as a "group" and tunes them separately with a slightly modified (English, Mullard) UHF television tuner. The TVRO signals are 36 MHz wide (and of course still FM) and we need to convert them again (in frequency) down to a low enough IF where they can be detected. Experimenter Steve Birkill has found that an English Mullard type ELC1043/05 UHF TV tuner makes a dandy tuneable second conversion system with only minor modifications. Unlike U.S. (of Canadian or Japanese, etc) UHF tuners designed for the American NTSC signals, the English (Mullard) tuner is capable of passing the full 36 MHz wide TVRO signal with only very minor modification. All American tuners checked have a 3 dB passband of not more than 10 to 11 MHz which simply means that they are not wide enough (even if modified) to handle the extra modulation/carrier width of the TVRO signal. Another advantage of the (European) UHF tuner, in this application, is that it has 20 dB of RF (500-1,000 MHz) gain and a quite respectable noise figure of under 5 dB. American market UHF tuners lack RF amplifier stages and consequently their front end noise figures are in the 12 dB and up region.

Birkill takes his low IF out at 35 MHz which allows him to use a Signetics 561 phase-locked-loop as a demodulator. Many of the commercial receivers also use phase-locked-loop demodulators, but Birkill's approach is unique since it allows the system user to change the effective bandwidth of the total system by varying the way the 561 is driven. This allows you to capture (that is, see) signals that are far weaker than would register on a standard 30- to 36-MHz wide IF satellite receiver; although admittedly the quality does suffer in the process. However, in his case, Steve Birkill has been able to produce very

continued on page 65



1

CIRCLE 106 ON FREE INFORMATION CARD

Realistic Model STA-2200 Receiver

LEN FELDMAN

CONTRIBUTING HI-FI EDITOR

REALISTIC AUDIO PRODUCTS ARE SOLD exclusively by the thousands of Radio Shack stores located throughout the United States and through their mail-order catalogs. The company has its own engineering staff, located in Fort Worth, Texas, to design the products which are then manufactured for them overseas. Their most sophisticated stereo receiver to date is the *model STA-2200* shown in Fig. 1.

The gold-colored front panel has no conventional dial scale or pointer. Instead, a large, highly visible digital display indicates the frequencies selected. The display reads-out in 200-kHz increments for FM and in 10-kHz increments for AM and is associated with a true frequency-synthesis tuning scheme. The display also serves as a real-time digital clock, indicating the time of day (including "AM" or "PM" notations) when non-radio program sources are being heard, or even in the AM and FM mode when a CLOCK pushbutton located behind a cleverly arranged sliding door panel is depressed. Program functions, such as phono and auxiliary, are also designated by alphabetic notations in the display area, as in reception of stereo FM signals.

Sliding up the aforementioned panel (located at the left of the display) reveals additional buttons for setting the hour and the minute of the digital clock display, for dimming the intensity of the display, for turning off output-power and signal-strength LED displays and for activating the "station-memory" feature. When the MEMORY-SET button is pushed, you have five seconds in which to enter the tuned-to frequency of six AM or six FM stations (a total of twelve) by depressing one of six buttons located beneath the digital readout area. Just to the left of the sliding panel are tape monitor and dubbing lever switches for operation of two connected tape decks for dub-

bing from one to the other.

The upper right of the panel is equipped with several tuning option buttons. Pressing the auto-tuning UP or DOWN buttons makes the tuner scan in the desired direction until the next usable signal is reached. A second pair of buttons (equipped with arrows, like the first pair) can be used for manual tuning. (The tuner continues to move up or down in frequency so long as the button is held down.) Just below those, a SCAN button, when depressed, makes the tuner scan to each of the memorized station frequencies in sequence, stopping for five seconds at each station for auditioning. If you want to stop the scan sequence, press the HOLD button during the 5-second interval. A POWER on/off button is also located at this end of the panel.

The center of the lower section of the front panel contains two banks of LED's (10 LED's per bank), which serve as power-output meters. These LED's are calibrated in watts, referred to 8-ohm loads. Calibration is either 100 watts or 10 watts per channel maximum, depending upon the setting of a pushbutton. Additional pushbuttons in that cluster include a mute button for FM, A and B speaker selector buttons, a tone control defeat button, mono/stereo switch, an MPX high-blend filter switch, and a loudness switch.

Concentric volume and balance controls are located at the lower right corner of the panel, as is the usual headphone connection jack. To the left of the power LED display are bass and treble tone controls, and between them are a pair of pushbutton switches that determine the frequency at which the tone controls begin to boost or cut (150 Hz or 300 Hz for the bass, 3 kHz or 6 kHz for the treble). A program selector switch is located at the extreme left of the panel and, in addition to its expected settings (phono, AM, FM and aux), also has a

RADIO-ELECTRONICS AUDIO LAB

R.E.A.L. SOUND

RATES

REALISTIC STA-2200

VERY GOOD

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setting for Dolby FM. The receiver is equipped with a complete Dolby FM decoder circuit.

The rear panel of the STA-2200 is shown in Fig. 2. A line fuseholder and switched and unswitched AC receptacles are at the left. Nearby are spring-loaded color-coded speaker terminals for the two pairs of speakers. There are also RCA-type phono jacks for use when your speakers are equipped with permanent cables terminating in matching phono plugs. Preamp-out/main-amp-in jacks are interconnected by a pair of removeable wire jumpers.

Two sets of tape-out/tape-in jacks are augmented by European type DIN multi-contact connectors. Other inputs (auxiliary and magnetic phono) are at the lower right of the panel, far removed from any AC hum fields, and a



2

MANUFACTURER'S PUBLISHED SPECIFICATIONS:

FM TUNER SECTION:

Usable Sensitivity: mono, 1.8 μ V (10.3 dBf). **Signal-to-Noise Ratio:** mono, 68 dB. **Image Rejection:** 75 dB. **Capture Ratio:** 1.5 dB. **IF Rejection:** 95 dB. **AM Suppression:** 55 dB. **Harmonic Distortion:** mono, 0.2%; stereo, 0.3%. **Stereo Separation, 1 kHz:** 48 dB.

AM TUNER SECTION:

Sensitivity: 10 μ V. **Distortion:** 1.0%. **Image Rejection:** 45 dB. **IF Rejection:** 47 dB. **AM Fidelity:** 40 Hz to 3,000 Hz, ± 6 dB.

AMPLIFIER SECTION:

Power Output: 60 watts per channel into 8 ohms, 20 Hz to 20 kHz. **Rated Harmonic Distortion:** 0.02%. **Input Sensitivity, for full output:** phono, 2.2 mV; high level, 160 mV. **Phono Overload:** 200 mV. **Signal-to-Noise Ratio:** phono, 65 dB; high level, 75 dB. **Bass Control Range:** ± 10 dB at 50 Hz or 100 Hz. **Treble Control Range:** ± 10 dB at 10 kHz or 20 kHz.

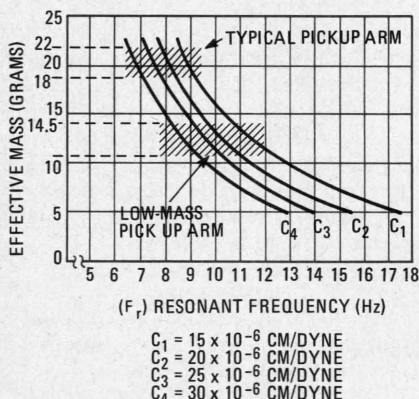
chassis ground terminal is located near the phono input jacks. Antenna screw terminals for AM, 75-ohm and 300-ohm FM antenna connections are provided along with a pivotable AM loopstick ferrite bar antenna.

Construction and circuit highlights

An internal view of the STA-2200 chassis is shown in Fig. 3. The FM front-end uses a dual-gate MOS FET and is tuned by variable-capacitance diodes, controlled by a microprocessor. The FM IF section uses 3 IC's and three

between 7.6 Hz and 12 Hz, thanks to its lower effective mass.

In addition to its lower mass, the Thorens Isotrack arm has a low 25-milligram (maximum) pivot friction in both the vertical and horizontal planes. Anti-skating is applied through a magnet arrangement that has the advantage over the familiar spring or weight arrangement that tends to introduce additional and variable fric-



tion components as the arm traverses the record.

According to Thorens, there is one additional advantage to be gained through the use of a low-mass arm. That is, lower susceptibility to vibration and shock from outside sources.

There is one other important consideration that should be discussed in any analysis of pickup arm design. That is the configuration of the bearing assembly and the length of the arm itself. Because so many of the records we play are warped, the bearing pivot point should ideally lie in the same plane as the playing surface. With such positioning, the longitudinal displacements caused by record warp are kept to a minimum, and the resulting wow-and-flutter components are lower. We are not discussing the wow-and-flutter that is inherent in the turntable drive system itself (that will be discussed next month). Rather, we are discussing that wow-and-flutter component that is generated entirely by vertical motion of the cartridge stylus as it is lifted and lowered by a warped record. Figure 9 illustrates this effect as it occurs with a pickup arm of normal length. The closer the pivot point is to the level of the record surface, the lower the wow and flutter produced by warped records. Extremely high levels of wow-and-flutter occur with short-arm designs that have a relatively high bearing-pivot point. Some of the so-called tangentially tracking arms that are available today have short arms, simply because tracking-angle error (the main reason for attempting a tangential-tracking arm in the first place) is eliminated in such designs. (The arm and cartridge cantilever assembly are always tangent to the groove being played). After investigating this approach, Thor-

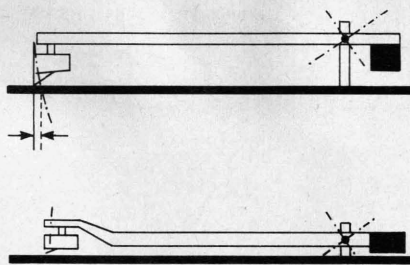


FIG. 9—WOW AND FLUTTER caused by record warp depends on height of pivot above record. Bottom design shows lower pivot.

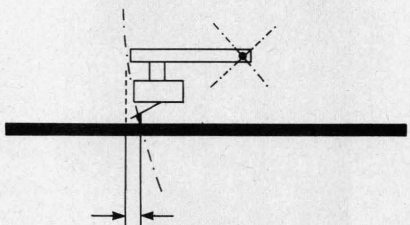


FIG. 10—RECORD-INDUCED WOW-AND-FLUTTER can be reduced in designs that have the arm pivot moved closer to the cartridge head.

ens concluded that while tangential arms are certainly feasible, the additional drive requirements for such arms make them too complicated to be practical.

In any case, referring once more to the wow-and-flutter problem inherent in short-arm designs (including the tangentially tracking ones currently available), consider the short, 4-cm arm constructed as shown in Fig. 10. The wow-and-flutter added by a warped record to that inherent in the turntable drive system itself works out to be:

TOTAL WARP EXCURSION	ADDED WOW-AND-FLUTTER
1 mm	0.12%
2 mm	0.24%
3 mm	0.36%

If we assume that a high-quality turntable exhibits its own basic wow and flutter component of around 0.05%, then with a 1 millimeter record warp the total wow-and-flutter will rise to 0.17%. A quick survey of your own records will show that you have very few that have less than 1 millimeter of warpage!

It is clear that the design of a complete turntable system involves many diverse factors, not the least of which is the fact that most phonograph records are less-than-perfect. To design a record playing system on the premise that records that will be played on it are perfectly flat, have completely concentric holes, and will be played in an environment that is totally free of outside vibration or noise is to ignore the real-world situation. In the next installment, we will talk about rumble, wow-and-flutter, mounting suspensions, and other factors relating to the turntable platter itself and its drive system. We will also try to show why one manufacturer's turntable, that has a -70 dB rumble figure, may, in fact, produce more rumble than another turntable that has a -50 dB rumble figure.

R-E

TV RECEIVERS

continued from page 59

high quality reception from the Russian stationary series of satellite transponders although he is working with signals 7-9 dB weaker than we have available here from North American domestic satellites, and he has acceptable (if not high quality) pictures from the much weaker Intelsat satellites (they run from 12 to 15 dB weaker than our domestic satellites). The balance of his receiver approach is pretty standard since once you have baseband video and audio there is only about one way to process it for conversion back to RF as an AM format signal.

To some people, being at baseband with the signal may seem like ending up at the wrong place. To view baseband directly, you feed the video and audio signals into a video "monitor" and a speaker. Not everyone has a video monitor of course and some means of getting the baseband signal back to a standard NTSC television channel (with the video portion amplitude modulated) is required.

A word about viewing the signal(s) at pure baseband; i.e. into and through a video monitor. This is the ultimate (high class) viewing technique since the baseband signals are of very high quality (48- to 54-dB signal to noise) and purity. However, this generally limits you to viewing the signals on a single monitor since video monitors tend to be expensive.

In Fig. 4 we have several methods suggested to get the baseband back to an RF channel. Clearly the baseband video and audio must be used to modulate a TV channel modulator device. Numerous circuits for these devices have appeared in **Radio-Electronics** through the years. One of the easiest ways to modulate back to RF is to use a LM1889 IC which is a complete (TV channel 3 or 4) RF carrier generator/modulator intended for TV games and home VTR's. If you already have a home VTR, you can simply loop the baseband video and audio to the home VTR's "camera" and "audio" inputs. This turns the VTR into a modulator for you and you can then watch the satellite TV signals on multiple TV receivers, connected to the VTR modulator through 75-ohm coaxial cable (as in a miniature TV distribution system).

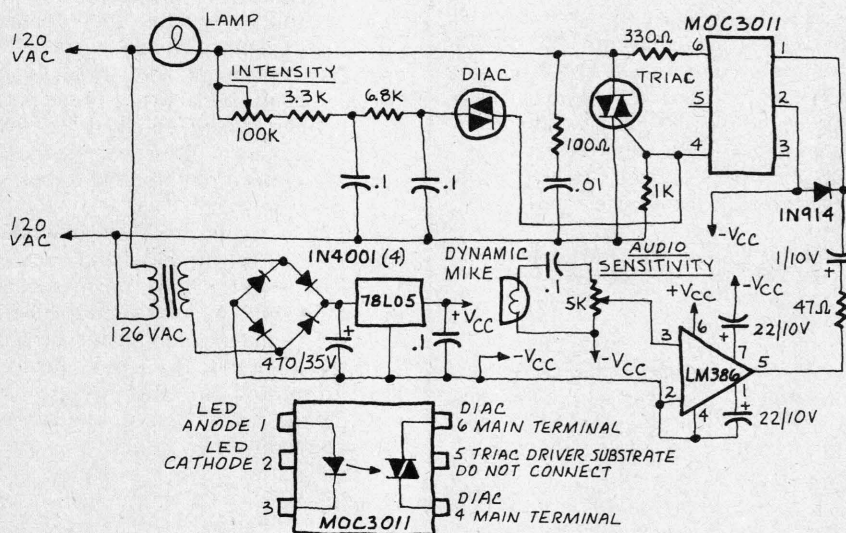
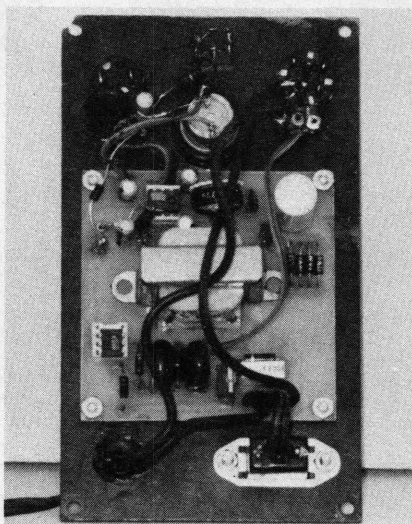
R-E

 **American Heart Association**
WE'RE FIGHTING FOR YOUR LIFE

SOUND-ACTIVATED LAMP DIMMER

MY NEW IDEA CONCERNS THE USE OF Motorola's MOC3011 opto-isolated triac driver. The triac driver consists of a LED and an optically coupled bilateral switch (better known as a DIAC). This device permits low-voltage control signals to control high-voltage, high-power loads. It essentially allows the user to create a solid-state relay for a fraction of the cost.

I wanted to make a simple and inexpensive device capable of sound modulating the intensity of a light source using a microphone built into the case. The device that I designed employs a conventional light dimmer circuit in addition to the sound activation circuit. The light dimmer may be used either independently of or in conjunction with the sound



circuit. When used separately, the AUDIO SENSITIVITY control is set to minimum and the light dimmer is used in the conventional manner. When the light dimmer is activated by sound the 100K pot sets the minimum intensity that will remain in the absence of sound and return to after it has been triggered to full intensity by the sound section. This feature is made possible by the MOC3011. The signal that controls the triac in the light dimmer is so low that it does not harm the bilateral switch in the MOC3011. Furthermore, when the bilateral switch in the MOC3011 triac driver is triggered by the LED, it causes the triac to conduct fully and override the preset light dimmer setting.

The construction of the control unit is simple and straightforward. If circuit as-

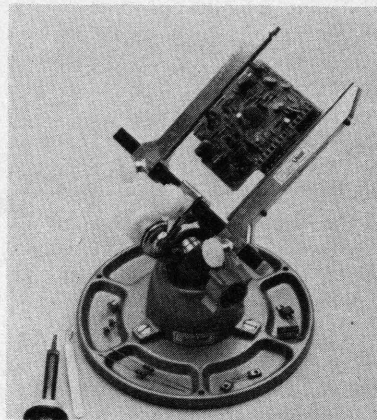
sembly is done using point-to-point wiring rather than a circuit board. Use **extreme caution** when wiring that portion of the circuit which is connected to the 120-volt power line. Be sure that during the wiring of the triac driver (MOC3011), no connection of any kind is made to pin 5. Set the microphone element into a hole drilled in the top of the case and secure it with a silicon rubber compound. The audio gain and light dimmer controls were placed near the microphone element to allow adequate separation between the low-voltage section and the 120-volt power control section. A chassis-mounted power socket was used as a convenient way to make the device flexible in use. If a high-power load is anticipated, be sure that adequate heat sinking is provided for the triac.

In addition to using this unit to produce scary effects for Halloween and cause your Christmas tree lights to dance with the Christmas music, you can also use this device to produce unique DISCO lighting effects with just about any lamp in the house. Several units placed around the room create a wild effect and there is no required connection to a sound source.—**David L. Holmes**

NEW IDEAS

This column is devoted to new ideas, circuits, device applications, construction techniques, helpful hints, etc.

All published entries, upon publication, will earn \$25 plus a Circuit Board Holder, Standard Base and Tray Base Mount from Panavise Products, Inc. (See photo below.) Selections will be made at the sole discretion of the editorial staff of **Radio-Electronics**.



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ED BATHGATE

THE MAJORITY OF PROBLEMS THAT OCCUR in a VCR are mechanical in nature. Problems caused by dirty heads, worn idlers, stretched belts, and jammed gears are perhaps most common, but VCR's also have their share of electrical problems. Such problems may be bad end sensors, burned out motors, power-supply problems, etc.

A good oscilloscope and a digital voltmeter can get you through the majority of VCR problems quickly and easily. However, problems involving the video heads, rotary transformer, head pre-amps, and head-switching circuits can be tough to troubleshoot. There are low-cost (\$60) video-head testers, but they won't indicate if a head is contaminated or if the gap is clogged; in either case the output will seriously be degraded.

You could replace the video head in question, but that requires that you have a spare head for every make and model of VCR you service. Changing heads is time consuming, and keeping lots of heads in stock is expensive. What's really needed is an instrument that can generate a known-to-be-good video-head playback signal, and one inexpensive source for such a signal is another VCR. A VCR creates that signal whenever it plays a tape, so a working VCR can be used to troubleshoot a broken VCR (see Fig. 1).

If you are repairing VCR's as part of a service business, you probably have more than one working VCR in the shop at any given time. What's needed is a video jumper cable to take the signal from the source VCR and inject it into the VCR being repaired. This project makes it possible to do just that, with no modifications to either VCR.

VCR operation

There are several signals that a video head generates during playback. The luminance and sync is a signal from 3.4 to 4.4 MHz, frequency-modulated by video luminance and sync information. The chroma, or color information, is a 629.371-kHz signal recorded by amplitude modulating the 3.4 MHz FM carrier. The

VCR HEAD AMP TESTER

This inexpensive piece of equipment can turn a second VCR into a valuable troubleshooting tool.

combined signals are usually referred to as video-head RF or RF envelope.

Two video heads are needed to "read" the information from a standard VHS videocassette (see Fig. 2). The two heads are mounted 180 degrees apart on a polished aluminum cylinder that spins counter-clockwise at 30 rpm. When one head completes a scan of the tape, the other head is ready to start its scan. In one scan, one video head generates a "field," a full top-to-bottom picture on the TV screen. The second video head also

generates a field, but it is interlaced with the field from the first head. The two interlaced fields make one frame.

A standard four-head VCR uses only two heads at a time, one pair for "SP" (two-hour standard play), and one pair for "EP" (six-hour extended play). If one of the video heads is bad, the VCR will send a full-size picture to the TV, but with only half the picture information, with every other field composed of "snow."

Each head has its own pre-amp, and the output of each one goes to an

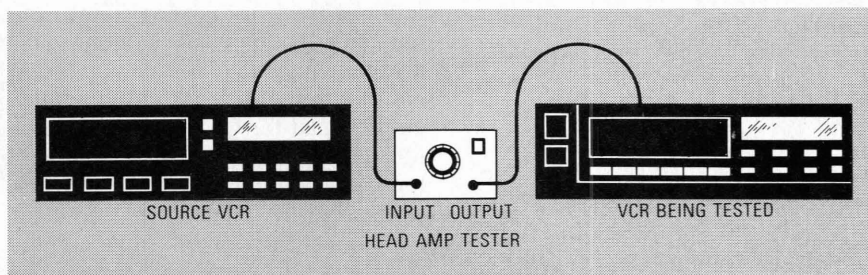


FIG. 1—THE VIDEO HEAD-AMP TESTER enables you to use a good signal from a working VCR to test a VCR with possible head problems.

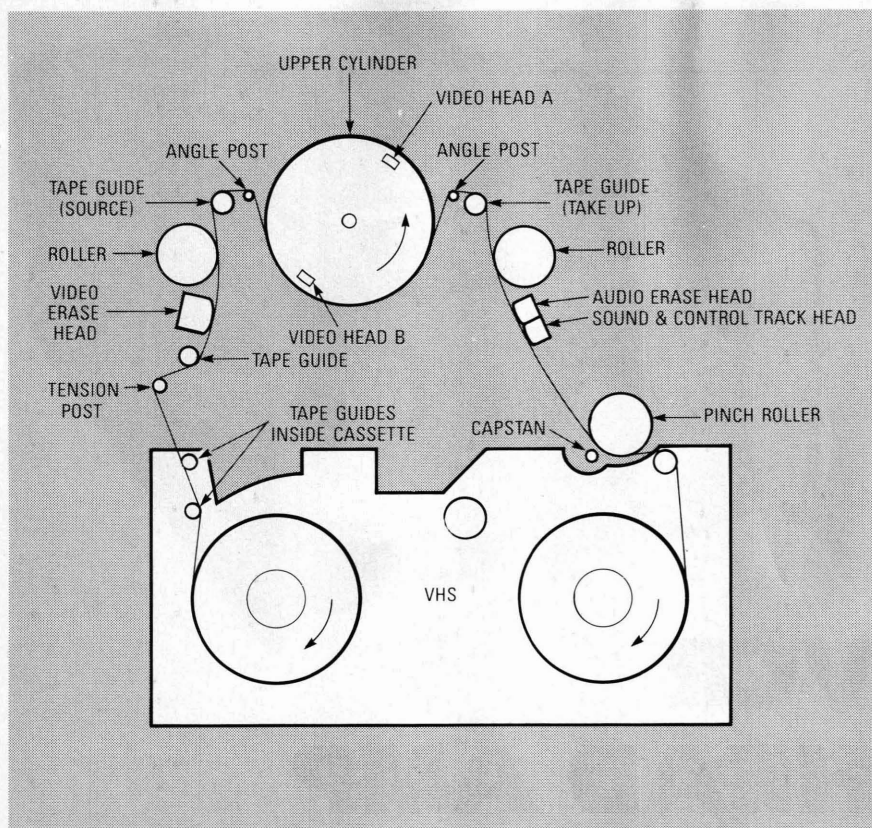


FIG. 2—VHS BASIC MECHANISM. Two video heads are needed to generate the standard VHS format. The two heads are mounted 180 degrees apart on a polished aluminum cylinder that spins counter-clockwise at 30 rpm.

electronic head switch (see Fig. 3). The head-switching circuit combines the outputs from each head pre-amp, by switching to the head which is in contact with the tape at that time. The head-switching control pulse is a 30-Hz square wave derived from the rotation of the head-cylinder motor. The output envelope (waveform *d*) is the

sum of the two individual head pre-amp envelopes (waveforms *a* and *b*).

If the head-switching pulse is not present, or if it's distorted or inverted in phase, the symptoms will be similar to bad heads or a bad pre-amp. Some examples of bad waveforms are shown in Fig. 4. Waveforms *a* to *d* are caused by mechanical misalignment of the tape guides, and the waveforms in *e* and *f* indicate proper alignment, but show a problem with the video heads, pre-amps, or head switcher.

Head-amp tester circuitry

The schematic for the tester is shown in Fig. 5. The input is an RF envelope from a working VCR, applied to Q1 through coupling-capacitor C1. Q1 is connected as an emitter follower, with a high-impedance input and a low-impedance output, and a voltage gain of 1.

Potentiometer R3 is used as the emitter load for Q1 and level control for the signal applied to Q2. Capacitor C2 is for improving the frequency response of R3. Transistor Q2 is also a 2N2222, wired in the same configuration as Q1, but with a lower output impedance in order to drive circuits in the VCR under test. The circuit draws

only 12 mA, so a 9-volt battery is well suited for the project.

Construction

The circuit should be built on a PC board, because RF as high as 4.5 MHz will be present. A single-sided board was used in the author's prototype with no problems. The board layout is very simple and can be drawn by hand directly on the copper with an etch-resist pen. See Fig. 6 for a parts-placement diagram; a foil pattern is provided in PC Service.

The assembled circuit should be mounted in a shielded box and coaxial leads should be used for input and output. Keep the lead length as short as possible (2-foot leads were used on the prototype with no problems).

Checkout

After assembly, check the voltages on Q1 and Q2, and the current draw, to verify proper circuit operation. Connect the VCR to be used as the signal

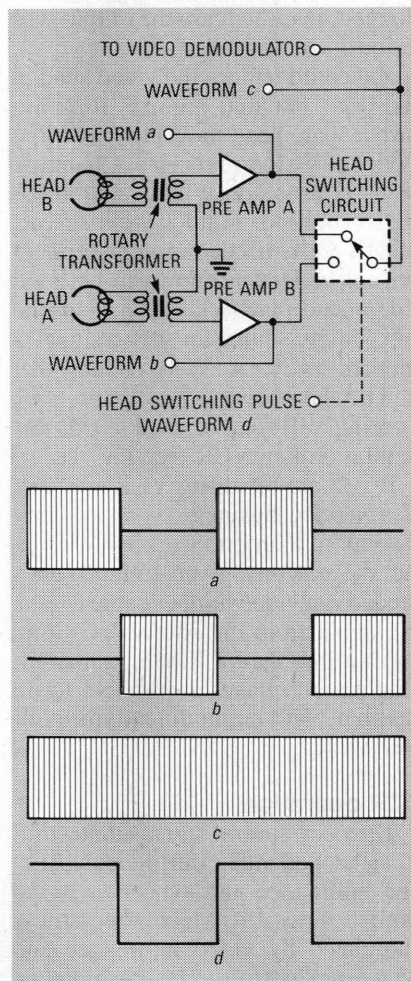


FIG. 3—EACH HEAD HAS ITS OWN PRE-amp, and the output of each one goes to an electronic switch that combines the outputs from each head pre-amp.

PARTS LIST

All resistors are 1/4-watt, 5%, unless otherwise indicated.

- R1, R4—100,000 ohms
- R2—220,000 ohms
- R3—10,000 ohms, audio-taper potentiometer
- R5—150,000 ohms
- R6—2200 ohms
- R7—1000 ohms

Capacitors

- C1, C3, C4—0.001 μ F, ceramic disc
- C2—39 pF, ceramic disc

Semiconductors

- LED1—red light-emitting diode
- Q1, Q2—2N2222 NPN transistor

Other components

- J1, J2—RCA-type jack
- S1—SPST on/off switch

Miscellaneous: Coaxial cable, PC board, metal case, solder, etc.

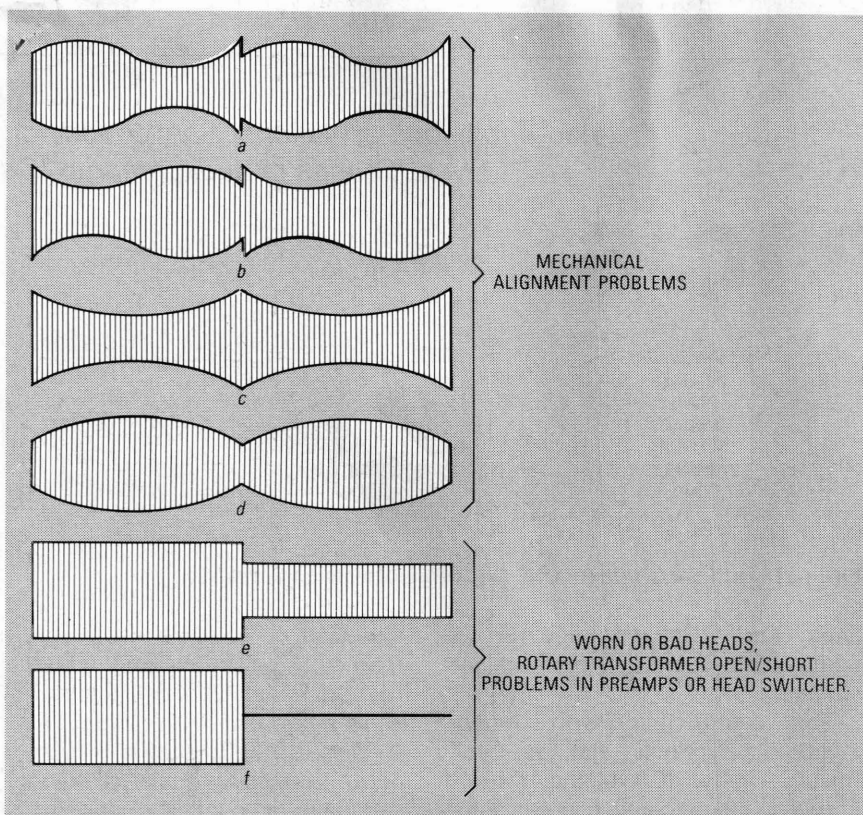


FIG. 4—IMPROPER WAVEFORMS. Waveforms a-d are caused by mechanical misalignment of the tape guides. The waveforms in e and f indicate proper alignment, but show that there's a problem with either the video heads, pre-amps, or head switcher.

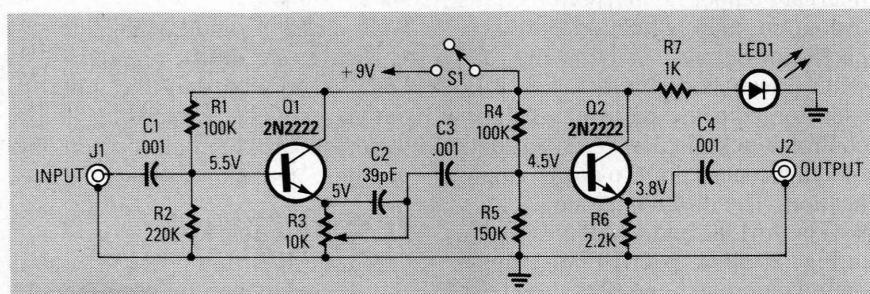


FIG. 5—THE SCHEMATIC for the head-amp tester.

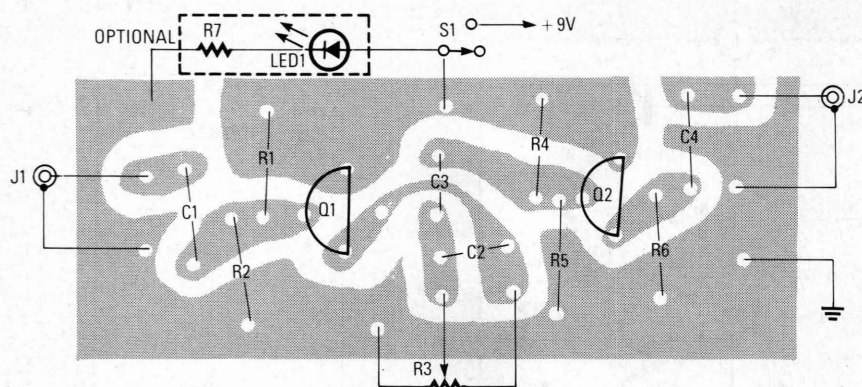


FIG. 6—PARTS-PLACEMENT DIAGRAM. Use the foil pattern provided in PC Service to make your own board.

source to a TV or monitor, and play a tape to use as the test signal; it can be a test pattern, or a home-made record-

ing of the news or some other show. Use an oscilloscope to check out the head RF envelope (Fig. 3-c), from the

source VCR for proper flatness. The RF envelope should be between 100- and 500-mV p-p in most VCR's.

Now turn on and connect the head-amp tester to the source VCR at the same point in the circuit that you measured the RF envelope (Fig. 3-c) with the oscilloscope. There may be a slight amount of signal degradation but if the entire picture disappears, it is loading down the source and the output signal will be unusable.

Check the output signal of the head-amp tester with the oscilloscope; it should be the same amplitude as the input signal with the level control at maximum. The output should be 0-V with the level control at minimum.

Using the tester

To substitute a signal in place of bad or questionable video heads, first put the source VCR into play, connect the head-amp tester, and adjust the output for 5-10-mV p-p. Put the VCR to be tested into play with a blank tape, and connect the output of the tester to the input of one of the head amps. That may be done at the connector end of the cable between the rotary transformer and the head amps. You can also capacitively inject the signal by clipping the output lead over the insulation of a non-shielded wire (no electrical connection), and increasing the output level to about $\frac{1}{2}$ to $\frac{3}{4}$ of maximum. Signals can also be injected into the input and output of the head switcher. The output level should be high and direct electrical connections should be made.

The rotary transformer (one that can couple a signal from a rotating drum to the rest of the circuitry) can be tested with the VCR under test in "stop" mode, but the source VCR must be in "play" to supply a signal. Connect the output lead directly across one head at a time, and measure the output at the pre-amp input connector. You should disconnect the pre-amp connector from the pre-amps if possible. The signal from the rotary transformer should be equal or greater in voltage than the applied signal voltage. Test each head and the corresponding transformer winding.

The head-amp tester is not going to replace any major test equipment, but it does help you to troubleshoot some problems. And, after all, why wouldn't you want all the help you can get?

R-E

AUDIO PRE AMP IC'S

An in-depth look at National Semiconductor's LM38X series of audio preamp IC's.

RAY M. MARSTON

ONE OF THE MOST VERSATILE SERIES OF linear preamp IC's is the LM38X versions from National Semiconductor. They're extremely useful for audio and tone-control applications, and have excellent ripple rejection, low signal distortion, wide bandwidth, and low noise. You'll find them in virtually any modern piece of audio gear. This discussion will investigate how they work, and look at several useful applications.

The LM38X IC's

Figure 1 shows a representative block diagram of a conventional stereo system channel with both volume and tone control. National Semiconductor produces five low-noise dual preamps in the LM38X series, the LM381, LM381A, LM382, LM387, and LM387A; the "A" denotes versions with superior noise figure performance. Figures 2-4 show the configurations of the three different versions for one of the two amplifiers in

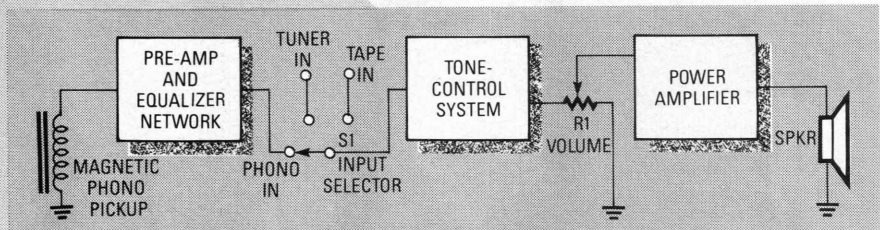


FIG. 1—BLOCK DIAGRAM OF ONE CHANNEL of a stereo system.

each Dual-In-Line Package (DIP), while Table 1 gives a performance summary.

Tone control may involve refinements like "scratch" and "rumble" filters. All five IC's in the LM38X family use single-ended power supplies, and have the same basic amplifier circuitry, but differ in internal details and pinouts. Also, all five have internal compensation, power-supply decoupling and regulation, large capacity for output-voltage swing, and wide power bandwidth. They'd be used for both the preamp and tone-control blocks in Fig. 1, since both functions occur prior to power amplification. The differences are:

- The LM381 and LM381A, shown in Fig. 2, allow external noise figure optimization and compensation (nar-

row-band or low-gain use). They're normally used differentially, but can be used single-ended for ultra-low-noise purposes.

- The LM382, shown in Fig. 3, doesn't provide for external compensation or single-ended operation, but has a built-in resistor matrix to let the user select from among several closed-loop gain and frequency-response options.

- The LM387 and LM387A, shown in Fig. 4, are utility versions of the LM381/1A, with only the input and output terminals accessible, and no provision for external compensation or single-ended operation.

LM381/1A basics

All the IC's in the LM38X family can be understood by examining the

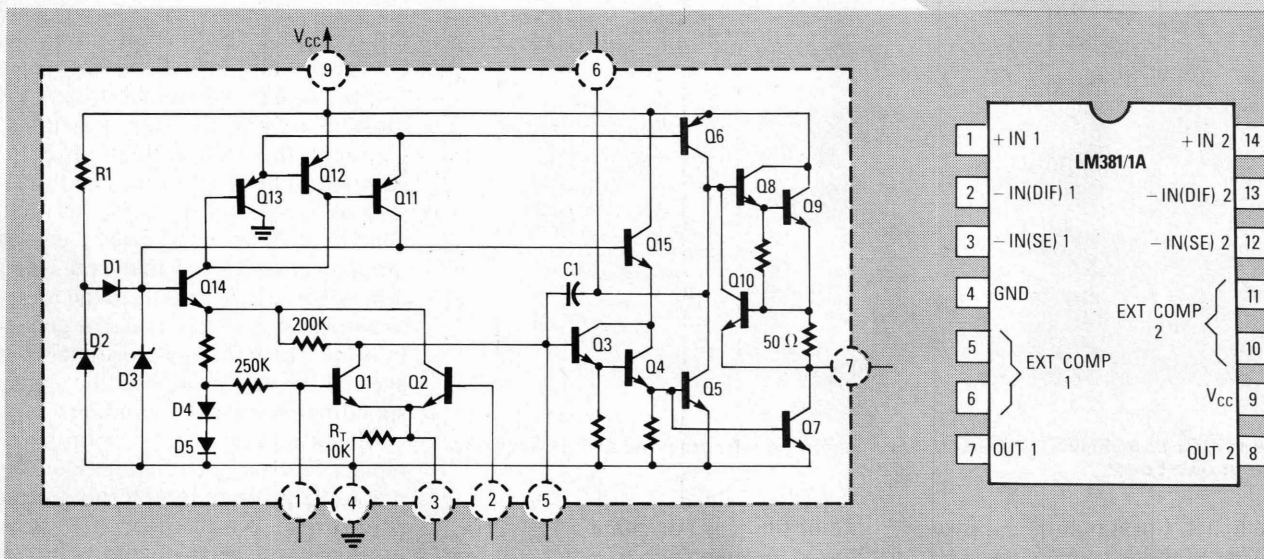


FIG. 2—THE LM381/1A DUAL LOW-NOISE PREAMP.

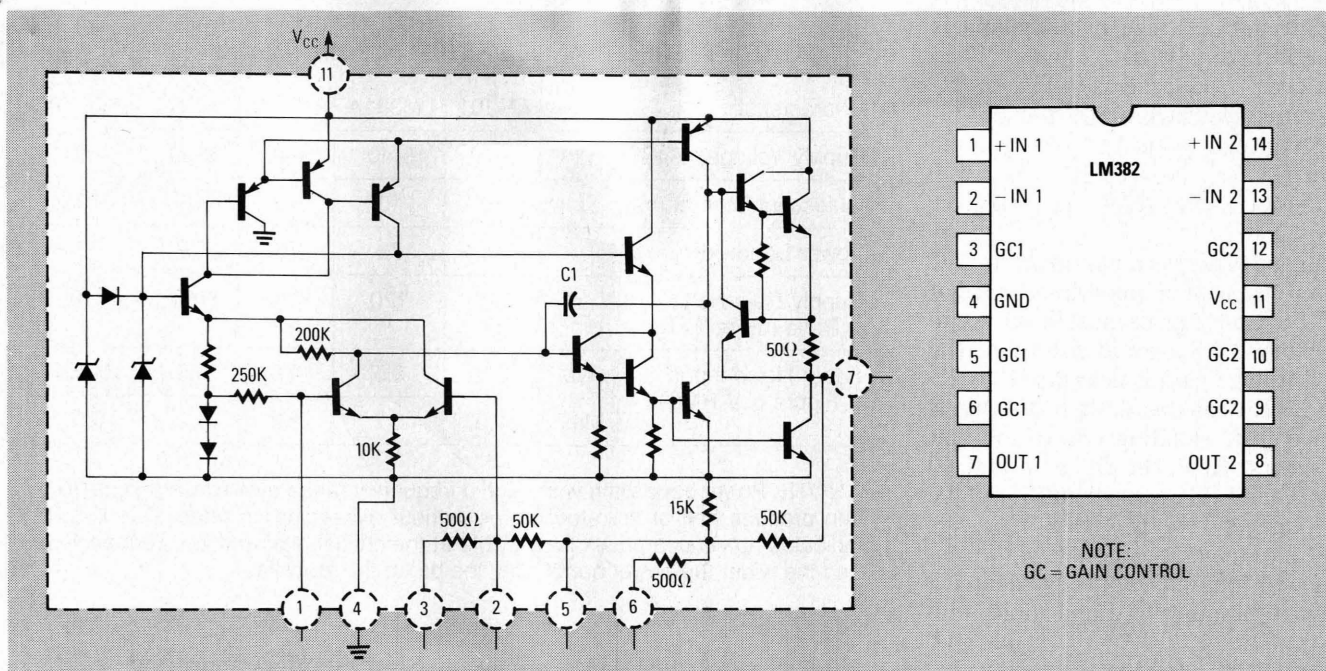


FIG. 3—THE LM382 DUAL LOW-NOISE PREAMP.

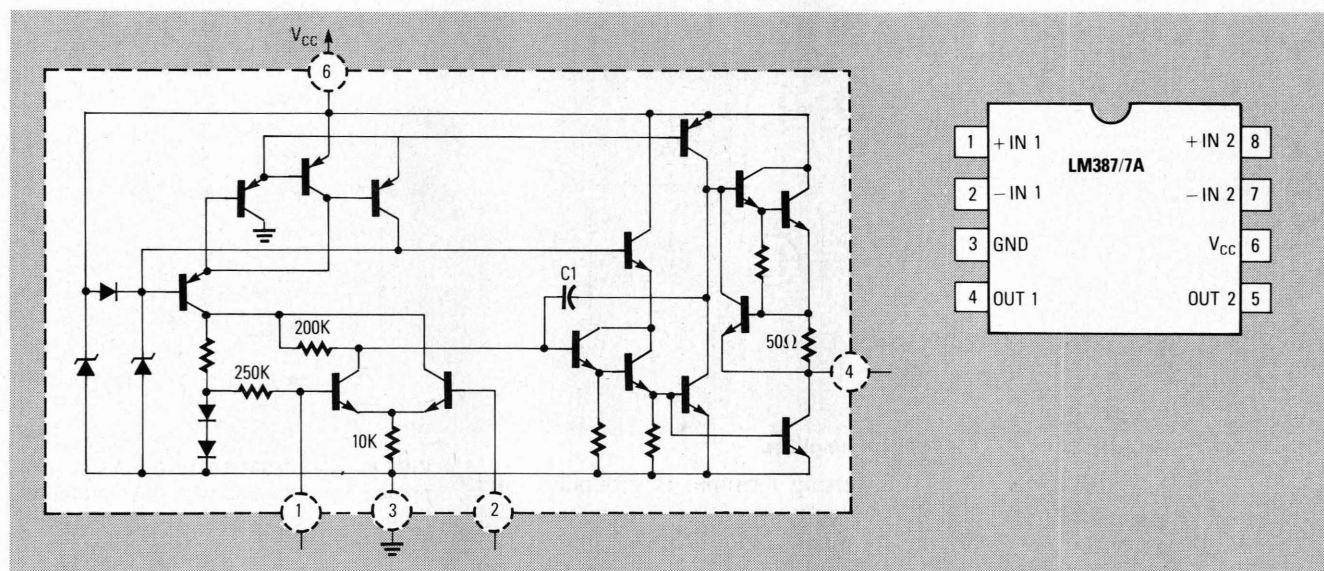


FIG. 4—THE LM387/7A DUAL LOW-NOISE PREAMP.

LM381/1A shown in Fig. 2. It has a first stage (Q1 and Q2), second stage (Q3–Q6), output stage (Q7–Q10), and bias network (Q11–Q15); Fig. 5 shows a simplified equivalent. The first stage is biased at 1.2 volts via R1, and can be operated either differentially or single-ended, although differential operation generates 41% more noise.

In differential use, the first stage has to be balanced by externally biasing the base of Q2 at 1.2 volts. In single-ended mode, Q2 has to be turned off by grounding its base, and Q1 has to be balanced by externally biasing the emitter of Q2 at 600 milli-

volts. The first stage has a differential voltage gain of 80, or 160 when used single-ended.

The second stage uses common-emitter Q5 and constant-current load Q6, and is driven by Q1 via Darlington emitter follower Q3–Q4. Its voltage gain is 2000, and it's internally compensated via C1 for unity gain at 15 MHz, giving stability at closed-loop gains of 10 or more. At lower gains, an external capacitor can go in parallel with C1 for compensation purposes.

The output stage uses Darlington emitter follower Q8–Q9, with active current sink Q7. Then, Q10 provides

short-circuit protection by limiting output current to 12 mA. The bias network gives 120 dB of supply-signal rejection, and includes the high-impedance constant-current generator Q11–Q12–Q13, which generates ripple-free reference voltage across D3. That reference voltage operates the first two stages via Q14 and Q15, and biases the base of Q1 internally.

Differential operation

In differential mode, the IC output is given a positive quiescent value independent of supply-voltage variations, by connecting divider R1–R2 as a DC negative-feedback loop, as

shown in Fig. 6. The inverting input is biased internally at 1.2 volts. When R1 and R2 are used as in Fig. 6, DC negative feedback makes the non-inverting input go to 1.2 volts, and the amplifier output to $1.2 \text{ volts} \times [1 + (R1/R2)]$. In practice: $R2 < 250K$.

Figure 7 shows a non-inverting AC amplifier with an input impedance of 250K; input signals must be limited to 300 mV RMS to avoid distortion. The DC voltage gain is determined by R1 and R2, while the desired AC gain is set by AC shunting one of the bias resistors. Here, the DC gain is fixed by R1 and R2 at less than 10, but the AC gain is fixed by R1 and R3 at 100.

That shunting technique can be expanded for frequency-dependent AC gain in various filter applications. Fig-

TABLE 1—PERFORMANCE OF THE LM381/1A/2/7/7A LINEAR IC'S

Characteristic	LM381	LM381A	LM382	LM387	LM387A
Supply Voltage (VDC)	9-40	9-40	9-40	9-30	9-40
Quiescent Current (mA)	10	10	10	10	10
Power Bandwidth (kHz)*	75	75	75	75	75
Supply Rejection Ratio (dB at 1 kHz)	120	120	120	110	110
Equiv. Noise Input Figure ($\mu\text{V RMS}$)	Typ.	0.5	0.5	0.8	0.65
	Max.	1.0	0.7	1.2	0.9

* NOTE: Power bandwidth is the audio frequency range over which an amplifier can produce half of its rated power, without exceeding its rated distortion. It indicates how much power is available at the critical high and low frequencies, and the wider the power bandwidth, the better the amplifier.

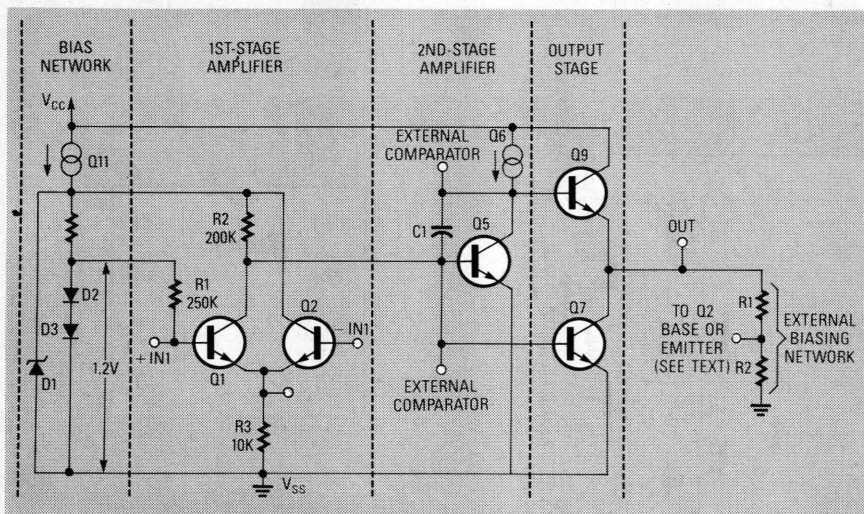


FIG. 5—EQUIVALENT CIRCUIT OF THE LM381/1A amplifier.

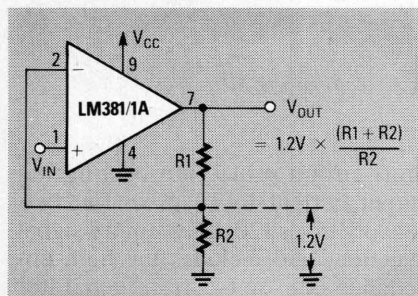


FIG. 6—DIFFERENTIAL BIASING OF THE LM381/1A.

ure 8 shows the same amplifier configuration used as a low-noise phono preamp with Recording Industries Association of America (RIAA) equalization, while Fig. 9 shows a similar tape-playback amplifier with National Association of Broadcasters (NAB) equalization. Figure 10 shows an inverting AC amplifier; here, the non-

inverting terminal is grounded, and the input signal is fed to the inverting terminal via R1. The AC gain is $R3/R2 = 10$, the quiescent output is +12 volts, and the input impedance is about R1. Figure 11 shows a unity-gain, 4-input audio mixer.

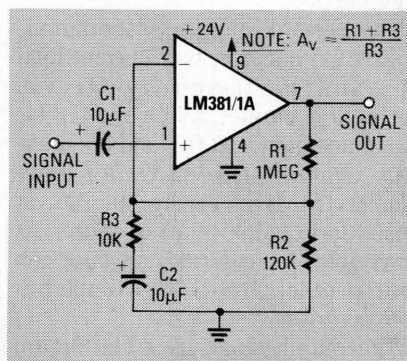


FIG. 7—A LOW-NOISE LM381/1A non-inverting amplifier, with a gain of 100.

Single-ended operation

The differential first stage of an LM381 composed of Q1-Q2 is powered via the internal 5.6-volt regulator, and the collector of Q1 is fed to

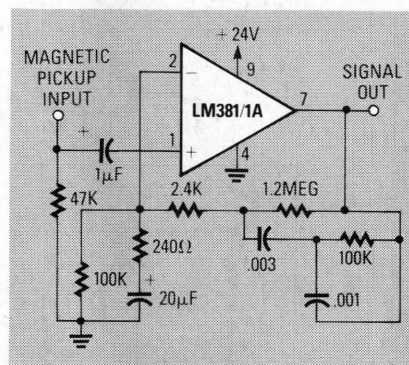


FIG. 8—AN LM381/1A USED AS A low-noise phono preamp with RIAA equalization.

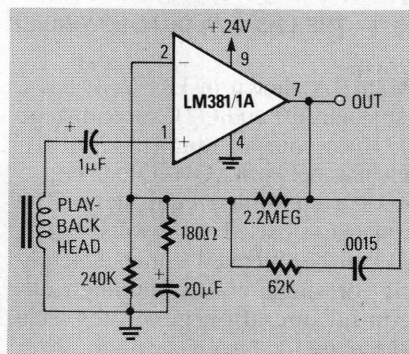


FIG. 9—AN LM381/1A used as a tape playback amplifier with NAB equalization.

the output via a DC amplifier. The IC can be operated in single-ended mode by grounding the base of (and disabling) Q2, but it needs to be biased

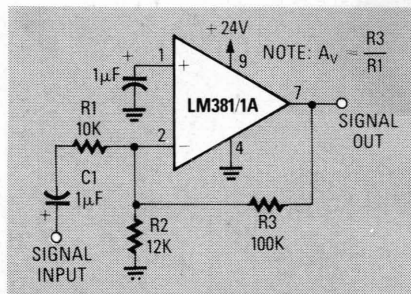


FIG. 10—THIS LM381A low-distortion (less than 0.05%) inverting amplifier has a gain of 10.

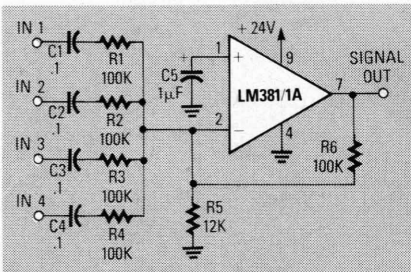


FIG. 11—THE LM381/1A is used here as a 4-input unity-gain audio mixer.

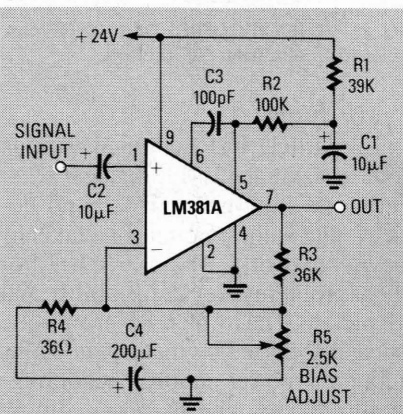


FIG. 12—THIS LM381A ultra-low-noise preamp has a gain of 1000.

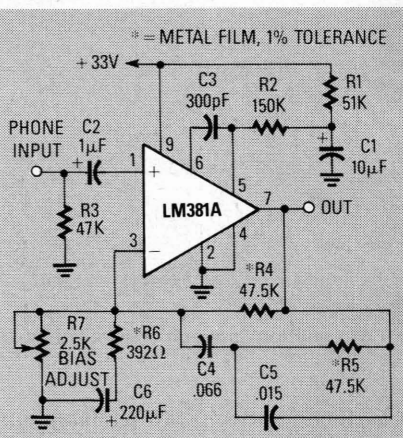


FIG. 13—LM381A ULTRA-LOW-NOISE magnetic phono preamp that includes RIAA equalization.

using emitter feedback.

Suitable DC biasing is obtained by connecting a voltage divider that ap-

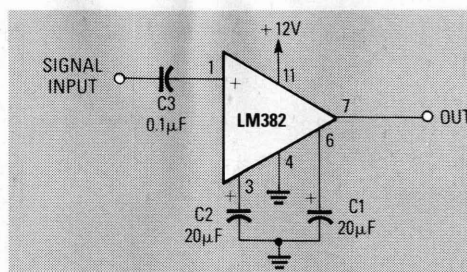


FIG. 14—AN LM382 FIXED-GAIN non-inverting amplifier with a 12-volt power supply.

GAIN	REQUIRED CAPACITOR
40dB	C1 ONLY
55dB	C2 ONLY
80dB	C1 AND C2

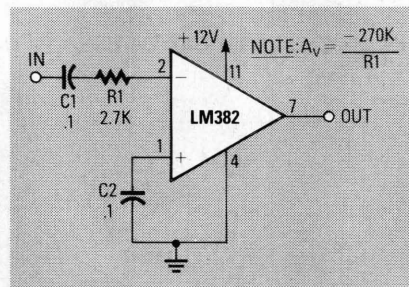


FIG. 15—THE LM382 is used here to make a 40-dB inverting amplifier.

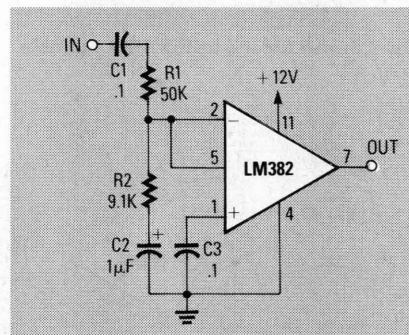


FIG. 16—SHOWN HERE IS AN LM382 unity-gain inverter.

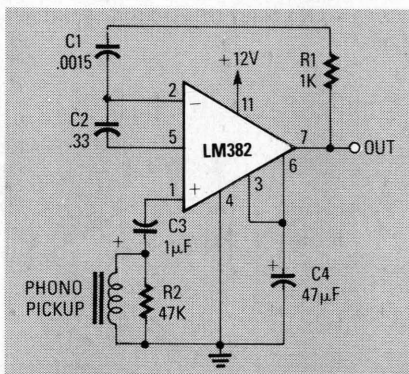


FIG. 17—LM382 phono preamp with RIAA equalization.

plies 600 millivolts to pin 3 when the IC output is at the desired DC level. If a quiescent +12-volt output is needed, the divider needs a DC voltage gain of 20.

In practice, the noise from the input transistor varies with collector current, and is minimized at 170 μA.

A single-ended LM381 is intended

for non-inverting use only, with a typical input impedance of 10K. Ideally, input signals should have source impedances below 2K, and all resistors should be of the low-noise, metal-film variety. Figure 12 is an ultra-low-noise version with a gain of 1000, where C3 limits the upper 3-dB frequency response to 10 kHz, and R5 adjusts the DC output voltage to half-supply value. Figure 13 is a magnetic phono preamp circuit that uses RIAA equalization, with the DC output voltage set by R7.

LM382 circuits

The internal circuitry of each half of a LM382 is identical to a LM381, except for a 5-resistor matrix and elimination of certain terminal connections. Eliminating those terminals means that an LM382 can't be used single-ended or externally compensated. The resistor matrix greatly simplifies bias- and filter-network design. The matrix is specifically intended for applications where the IC is powered from a +12-volt supply. Figures 14-17 show various ways to use the LM382 with a +12-volt supply. Figure 14 shows a non-inverting amplifier with 40, 55 or 80 dB of AC gain. Figure 15 shows an inverting amplifier with 40 dB gain, Fig. 16 shows a unity-gain inverting amplifier, and Fig. 17 shows a phono preamp with RIAA equalization.

LM387 circuits

The internal circuitry of each half of a LM387/7A is identical to an LM381, except for eliminating certain terminal connections, letting the IC be used differentially without external compensation. The IC is nevertheless quite versatile, and Figs. 18-24 show some practical applications. Figure 18 shows how to connect an LM387 as a non-inverting amplifier with an AC gain of 52 dB. The DC gain and quiescent output voltage of the amplifier circuit are determined

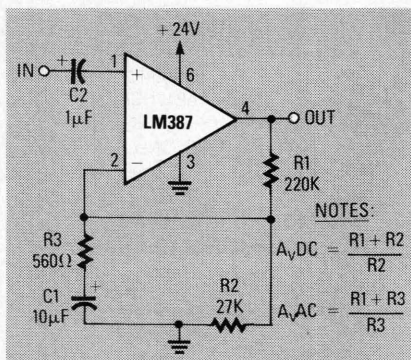


FIG. 18—LM387 NON-INVERTING AC amplifier with a gain of 52 dB.

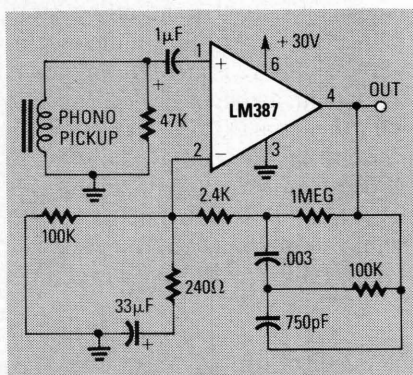


FIG. 19—LM387 PHONO PREAMP with RIAA equalization.

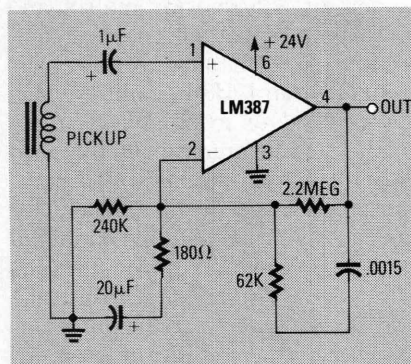


FIG. 20—AN LM387 USED AS a tape playback amplifier (NAB).

by R1 and R2, and the AC gain by R1 and R3. Figure 19 shows an LM387 used as a phono preamp with RIAA equalization, while Fig. 20 shows how it can be used as a NAB tape playback amplifier for use in all kinds of devices ranging from cassette players to telephone-answering machines.

Figure 21 shows an active tone control giving unity gain with its controls in the "flat" position, or 20 dB of boost or rejection with the controls fully rotated. The "rumble" filter of Fig. 22 is a 2nd-order high-pass active filter that rejects signals below 50 Hz at 12 dB/octave. Figures 23 and 24 show various ways of using an LM387 in inverting mode in active filters. The

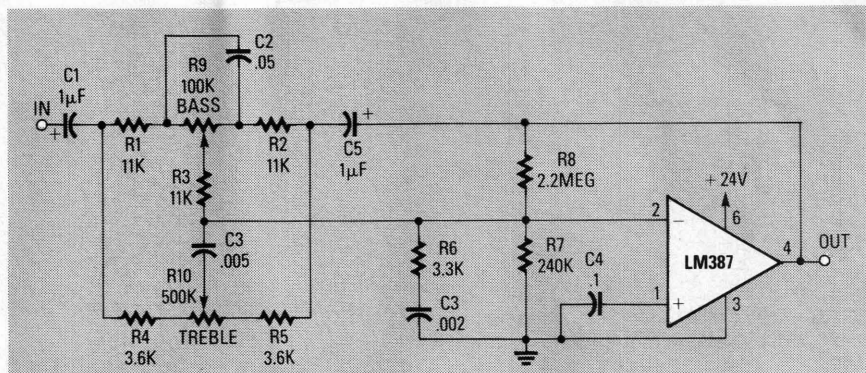


FIG. 21—THE LM387 CAN BE USED TO MAKE an active tone-control circuit.

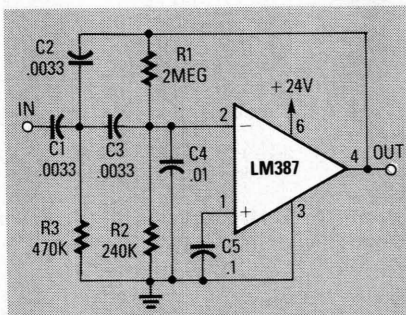


FIG. 22—AN LM387 USED AS A "rumble" filter.

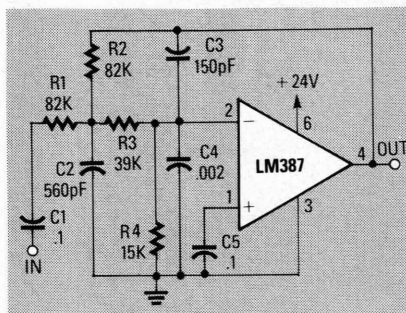


FIG. 23—AN LM387 USED AS A "scratch" filter.

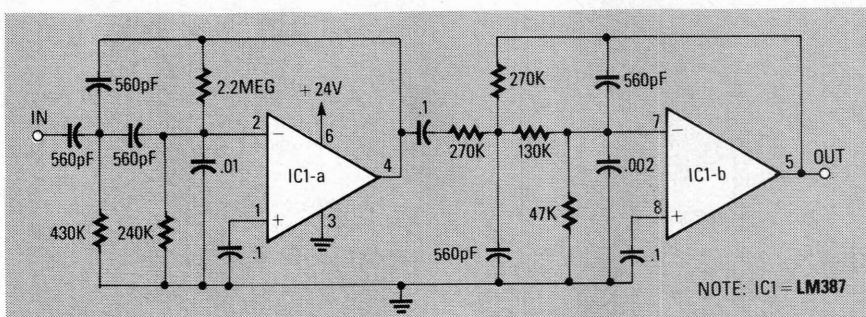


FIG. 24—BOTH HALVES OF AN LM387 make up a two-section "speech" filter (300 Hz–3 kHz).

"scratch" filter of Fig. 23 is a 2nd-order low-pass filter that rejects signals above 10 kHz. The "speech" filter of Fig. 24 consists of a 2nd-order high-pass and a 2nd-order low-pass filter in series, to give 12 dB/octave rejection to signals outside 300 Hz–3 kHz.

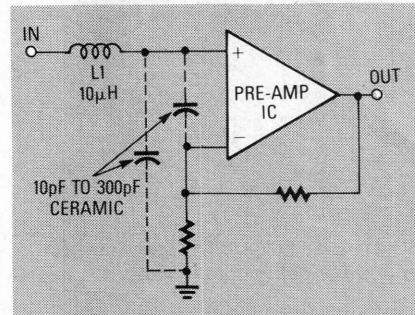


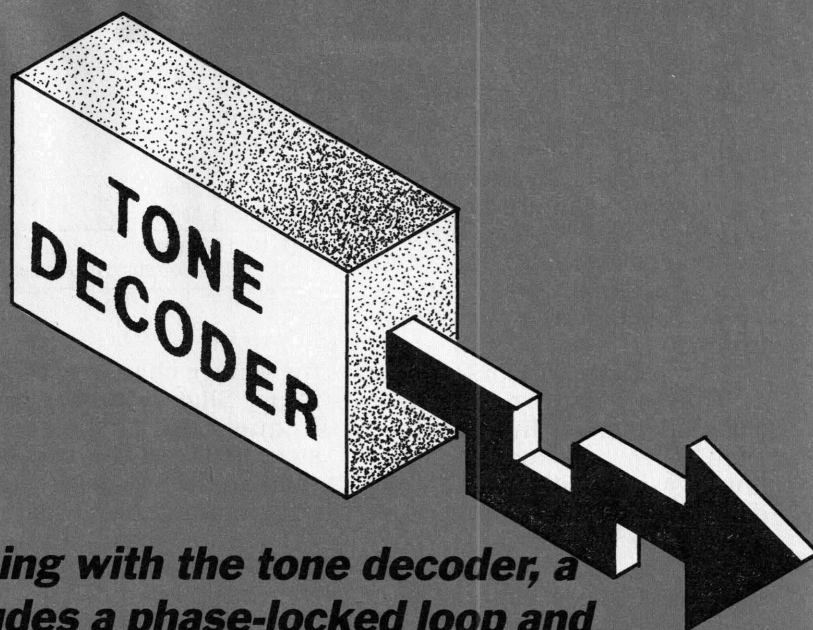
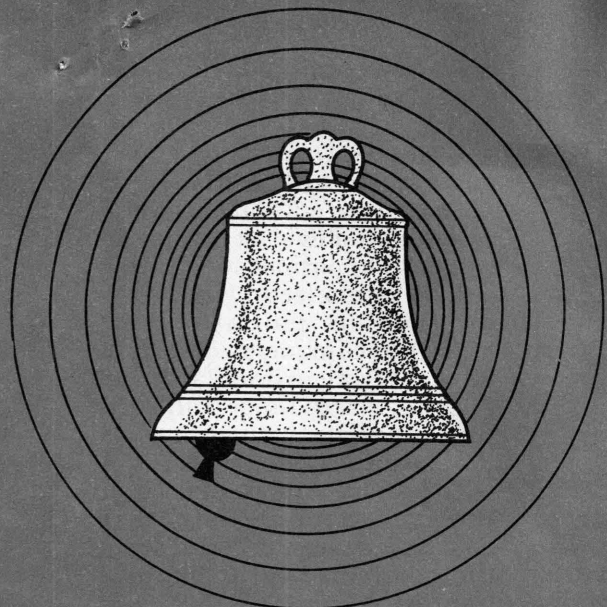
FIG. 25—THE CIRCUIT SHOWN HERE can be used to eliminate RF pickup.

Usage hints

This article has examined various circuits using the LM38X linear IC's. These are high-gain, wide-band devices, and some care must be taken if they're to work. The two most frequent problems are RF instability and RF pickup. The former, RF instability, is usually caused by inadequate high-frequency power supply decoupling. In all preamps, the IC power supply has to be RF-decoupled by wiring a

0.1 μF ceramic or 0.001 μF tantalum capacitor across the IC power pins. The latter, RF pickup, manifests itself as AM demodulation. It can usually be eliminated with a 10-μH RF choke in series with an IC's input terminals, or by also decoupling the input with a capacitor, as in Fig. 25.

R-E



Learn more about designing with the tone decoder, a versatile circuit that includes a phase-locked loop and responds to selected input frequencies

TONE DECODER

THIS ARTICLE, THE FOURTH IN A SERIES on phase-locked loops, is about a tone and frequency decoder monolithic integrated circuit. The tone decoder IC contains a stable phase-locked loop and a transistor switch that produces a grounded squarewave when a selected tone is introduced at its input. Tone decoders can decode tones at various frequencies. For example, it can detect telephone Touch Tones. The tone-decoder ICs are also found in communications pagers, frequency monitors and controllers, precision oscillators, and telemetry decoders.

The last three articles in this

RAY MARSTON

series explained the basic operating principles of the phase-locked loop and then went on to examine popular PLL ICs. Those included the Harris CD4046B PLL IC, the Philips (formerly Signetics) NE565 PLL IC, and the NE566 function generator IC.

This article is based on the Philips NE567 tone decoder/phase-locked loop. The device is a low-cost commercial version of the 567 packaged in an eight-pin plastic DIP. Figure 1 shows the pin configuration of that package, and Fig. 2 shows the

internal block diagram of the device. It can be seen that its principal blocks are the phase-locked loop, a quadrature phase detector, an amplifier, and an output transistor. The phase-locked loop block contains a current-controlled oscillator (CCO), a phase detector and a feedback filter.

The Philips NE567 has an operating temperature range of 0 to +70°F. Its electrical characteristics are nearly identical to those of the Philips SE567, which has an operating temperature range of -55 to +125°. However, the 567 has been accepted as an industry standard tone decoder, and it is alternate-

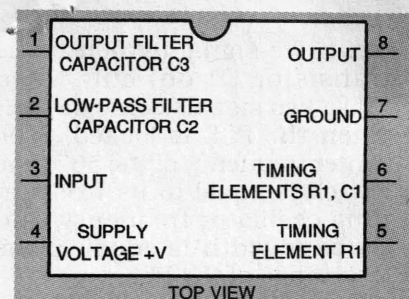


FIG. 1—PINOUT FOR AN NE567 TONE decoder in a eight-pin DIP package.

sourced by many other multinational semiconductor integrated circuit manufacturers.

For example, Analog Devices offers three versions of its AD567, Exar offers five versions of its XR567, and National Semiconductor offers three versions of its LM567. All of the different brands and models of this device will work in the circuits described in this article. Because

of the similarities between these devices, they will be referred to collectively as the "567" for the remainder of this tone decoder article.

The 567 basics

The 567 is primarily used as a low-voltage power switch that turns on whenever it receives a sustained input tone within a narrow range of selected frequency values. Stated in another way, the 567 can function as a precision tone-operated switch.

The versatile 567 can also function as either a variable waveform generator or as a conventional PLL circuit. When it is organized as a tone-operated switch, its detection center frequency can be set at any value from 0.1 to 500 kHz, and its detection bandwidth can be set at any value up to a maximum of 14% of its center frequency. Also, its output switching delay can be varied over a wide time range by the selection of external resistors and capacitors.

The current-controlled oscillator of the 567 can be varied over a wide frequency range with external resistor R1 and capacitor C1, but the oscillator can be controlled only over a very narrow range (a maximum of about 14% of the free-running value) by signals at pin 2. As a result, the PLL circuit can "lock" only to a very narrow range of preset input frequency values.

The 567's quadrature phase detector compares the relative frequencies and phases of the input signal and the oscillator output. It produces a valid out-

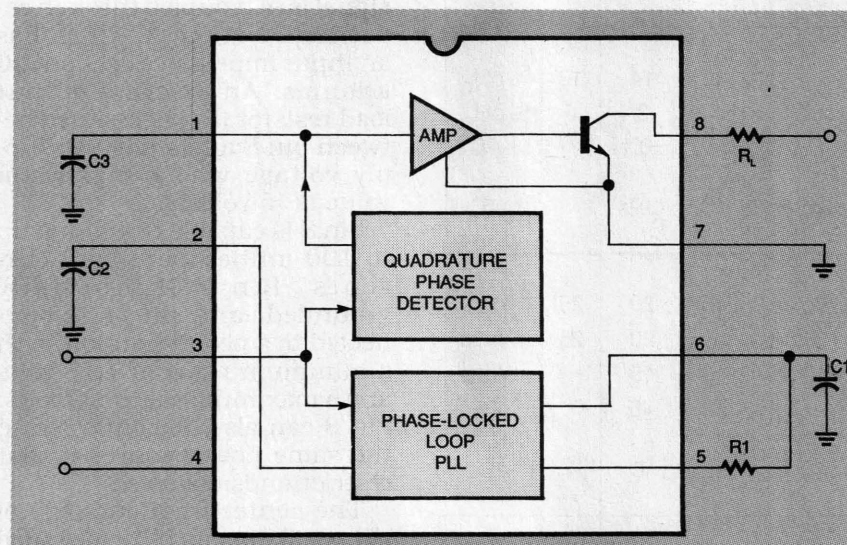


FIG. 2—BLOCK DIAGRAM OF AN NE567 TONE DECODER.

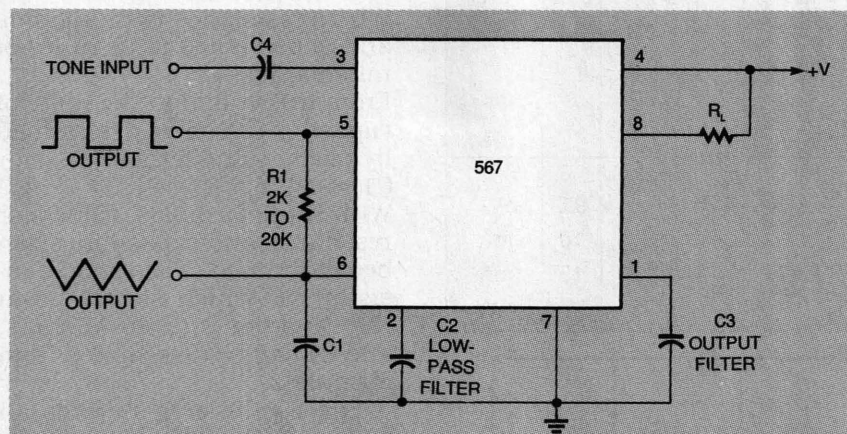


FIG. 3—TYPICAL CONNECTION DIAGRAM of a 567 tone decoder showing output waveforms at pins 5 and 6.

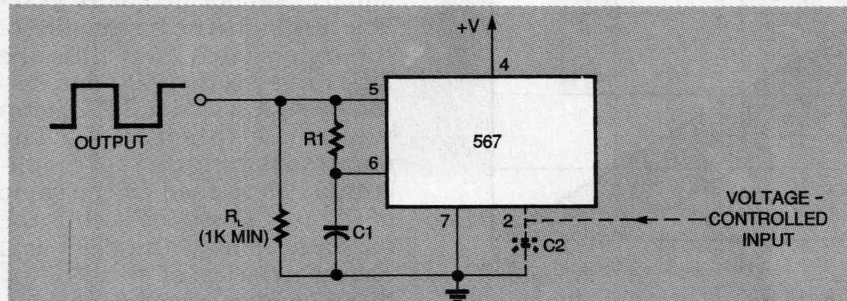


FIG. 4—PRECISION SQUAREWAVE generator based on the 567's 20-nanosecond rise and fall times.

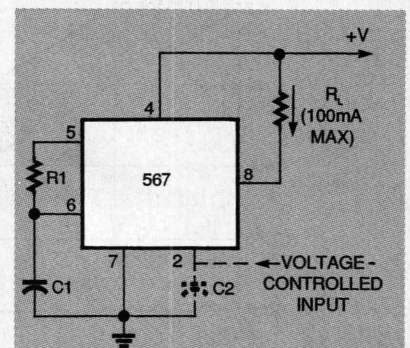


FIG. 5—PRECISION SQUAREWAVE generator based on a 567 configured for a high-current output.

TABLE 1—ELECTRICAL CHARACTERISTICS

PARAMETER	CONDITIONS	NE567			UNIT
		Min	Typ	Max	
CENTER OF FREQUENCY					
Highest center frequency (f_o)			500		kHz
Center frequency stability	-55 to +125°C		35±140		ppm/°C
	0 to +70°C		35±60		ppm/°C
Center frequency distribution		-10	0	+10	%
Center frequency shift with supply Voltage			0.7	2	%/V
DETECTION BANDWIDTH					
Largest detection bandwidth		10	14	18	% of f_o
Largest detection bandwidth skew			3	6	% of f_o
Largest detection bandwidth—variation with temperature			±0.1		%/°C
Largest detection bandwidth—variation with temperature	$V_i = 300\text{mVrms}$		±2		%/°C
INPUT					
Input resistance		15	20	25	kΩ
Smallest detectable input voltage	$I_L = 100\text{mA}$		20	25	mVrms
Largest no-output input voltage	$I_L = 100\text{mA}$	10	15		mVrms
Greatest simultaneous outband signal to inband signal ratio			+6		dB
Minimum input signal to wideband noise ratio	$B_n = 140\text{kHz}$		-6		dB
OUTPUT					
Fastest on-off cycling rate			$f_o/20$		
"1" output leakage current	$V_B = 15\text{V}$		25	0.01	μA
"0" output voltage	$I_L = 30\text{mA}$		0.2	0.4	V
	$I_L = 100\text{mA}$		0.6	1.0	V
Output fall time	$R_L = 50\Omega$		30		ns
Output rise time	$R_L = 50\Omega$		150		ns
GENERAL					
Operating voltage range		4.75		9.0	V
Supply current quiescent			7	10	mA
Supply current—activated	$R_L = 20\Omega$		12	15	mA
Quiescent power dissipation			35		mW

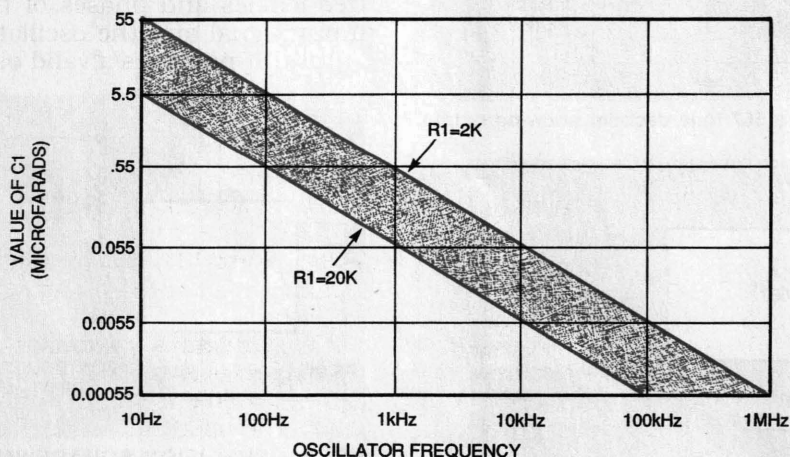


FIG. 6—RESISTOR-CAPACITOR SELECTION GUIDE for the current-controlled oscillator section of the tone decoder.

put-drive signal (which turns transistor Q1 on) only when these two signals coincide (i.e., when the PLL is locked). The center frequency of the 567 tone switch is equal to its free-running oscillator frequency, and its bandwidth is equal to the lock range of the PLL.

Figure 3 shows the basic connections for a 567 organized as a tone switch. The input tone signal is AC coupled through capacitor C4 to pin 3, which has an input impedance of about 20 kilohms. An external output load resistor (R_L) is inserted between pin 8 and a positive supply voltage whose maximum value is 15 volts.

Pin 8 is capable of sinking up to 100-milliamperes load currents. Pin 7 is normally grounded, and pin 4 is connected to a positive supply with a minimum value of 4.75 volts and a maximum value of 9 volts. Pin 8 can also be connected to the same power source if that restriction is observed.

The center frequency (f_o) of the oscillator can be determined by the formula:

$$f_o = 1.1/(R1 \times C1) \quad (1)$$

Where resistance is in kilohms and capacitance is in units of microfarads

From this equation the value of capacitor C1 can be determined by transposing terms:

$$C1 = 1.1/(f_o \times R1) \quad (2)$$

With these formulas, values for resistance and capacitance can be determined. The value of resistor R1, which should be in the range of 2 to 20 kilohms, and C1 can be determined from formula 2.

The oscillator generates an exponential sawtooth waveform that is available at pin 6 and a square waveform that is available at pin 5. The bandwidth of the tone switch (and thus the lock range of the PLL) is determined by C2 and a 3.9 kilohm resistor within the IC. The output switching delay of the circuit is determined by the value of C3 and a resistor within the IC. Table 1 lists the electrical characteristics of the Philips NE567 which has nearly identical characteristics to all other brands of the 567.

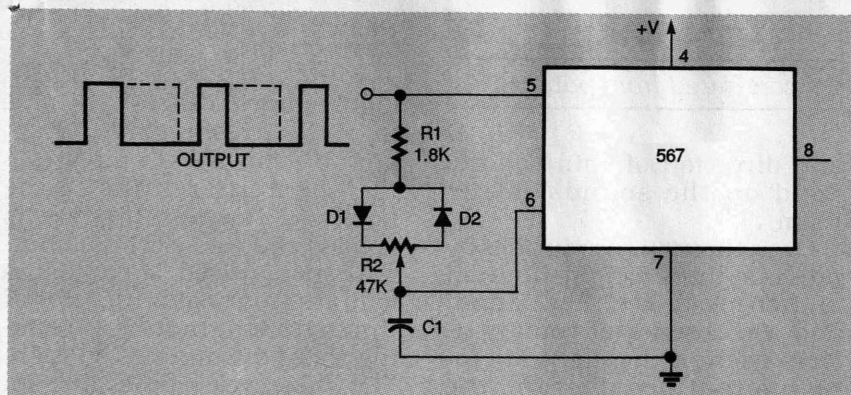


FIG. 7—SQUAREWAVE GENERATOR WITH A VARIABLE mark/space ratio output.

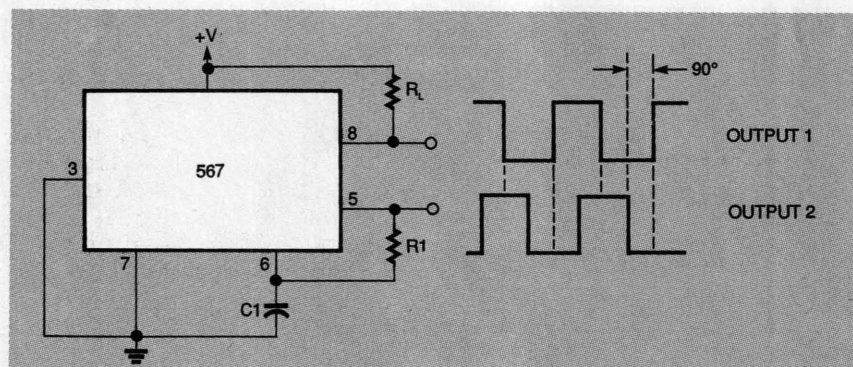


FIG. 8—SQUAREWAVE GENERATOR WITH quadrature outputs.

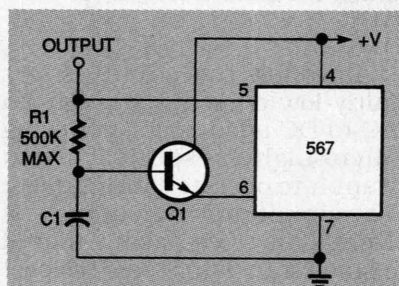


FIG. 9—TRANSISTOR BUFFER increases the permissible value of resistor R1.

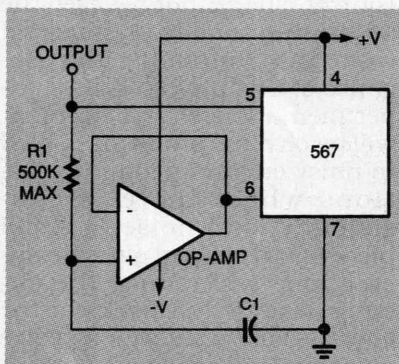


FIG. 10—OPERATIONAL AMPLIFIER buffer increases the permissible resistor value without waveform symmetry loss.

Oscillator design

Figures 4 and 5 illustrate how to obtain various precision

to-peak amplitude equal to the supply voltage value minus 1.4 volts. It can be externally loaded by any resistance value greater than 1 kilohm without adversely affecting the circuit's function. Alternatively, the squarewave output can be applied (in slightly degraded form) to a low impedance load (at peak currents up to 100 milliamperes at pin 8 output terminal, as shown in Fig. 5

By applying formulas 1 and 2 for oscillator frequency and capacitance, respectively, as presented earlier, various values can be determined. Again, R1 must be restricted to the 2 to 20 kilohm range. To save time in making this determination, component values as they relate to oscillator frequency can be read directly from the nomograph, Fig. 6.

For example, if you want the decoder's oscillator to operate a 10 kHz, the values for C1 and R1 could be either 0.055 micro-

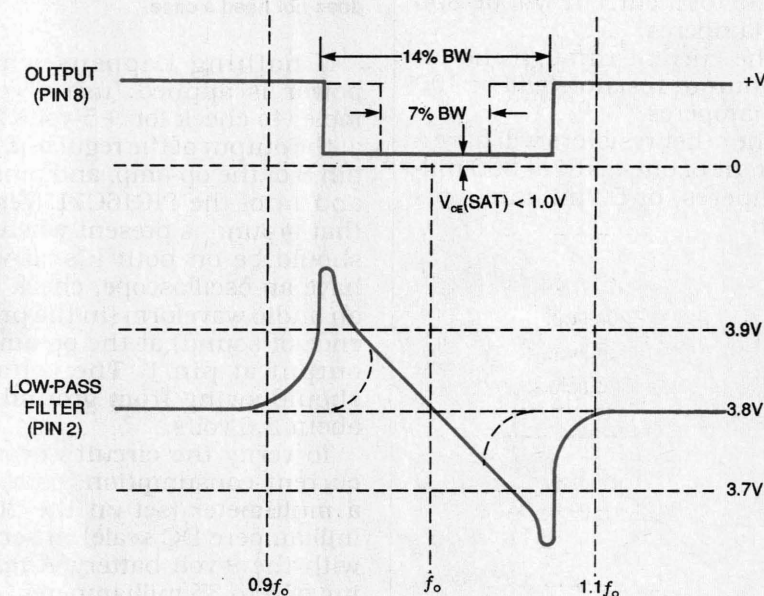


FIG. 11—WAVEFORMS AT PINS 2 AND 8 under in-band input voltage conditions.

squarewave outputs from the 567. The nonlinear ramp waveform available at pin 6 has only limited usefulness, but the squarewave available at pin 5 has excellent characteristics. As shown in Fig. 4, that output can have both 20-nanosecond rise and fall times.

This squarewave has a peak-

farads and 2 kilohms or 0.0055 microfarads and 20 kilohms, respectively.

The oscillator's frequency can be shifted over a narrow range of a few percent with a control voltage applied to pin 2 of the 567. If this voltage is applied, pin 2 should be decoupled by

(Continued on page 68)

RETRO REMOTE

continued from page 38

dress DIP switches open, address zero with all eight switches closed, or anything in between. Regardless of the address you select, be sure to set the same address on the receiver/decoder board.

Apply power to the receiver and connect a 9-volt battery to the transmitter. Test the training transmitter and receiver by aiming the transmitter at the receiver and pressing the transmit switch. If the circuit is working correctly, the valid transmission LED on the receiver will light up as long as you hold down the transmit switch. The ν LED should light regardless of the settings of the DATA DIP switches (S2 a-d). If the LED does not light, find and repair the mistake.

Follow the manufacturer's instructions for programming the learning remote. Operate the

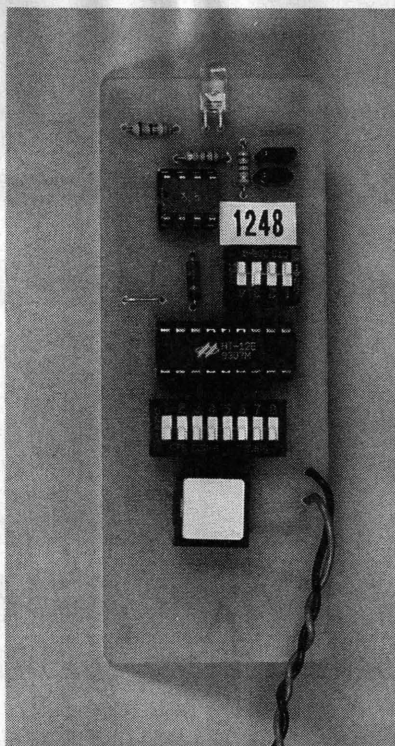


FIG. 7—COMPLETED TRANSMITTER BOARD. This board allows unique coding that won't interfere with nearby remote-controlled equipment.

training transmitter as you would any other remote control. As discussed earlier, the power-on command is decoded by the receiver as decimal 15. But, because the training transmitter understands only BCD, set all four data DIP switches at logic high (1111).

Now activate the learning remote's learning mode, press the on button, and press the transmit button on the training transmitter until the learning remote indicates that it has received the command. Next set the mute function as decimal 14 (1110), volume-up as decimal 13 (1101), and volume-down as 12 (1100).

How you program the remaining 12 receiver command codes is your choice. You might want to map 0 through 9 to buttons 0 through 9 on the remote. That still allows for two additional commands. Don't forget to program all your other remote controls into the learning remote too. Ω

TONE DECODER

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C2, whose value should be approximately double that of C1.

The circuits in Figs. 4 and 5 can be modified in several different ways, as shown in Figs. 7 to 10. In Fig. 7, the duty cycle or mark/space ratio of the generated waveform is fully variable over the range of 27:1 to 1:27 with trimmer potentiometer

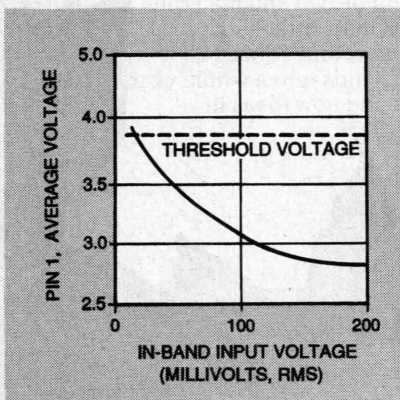


FIG. 12—TRANSFER FUNCTION of average voltage at pin 1 with respect to in-band input voltage.

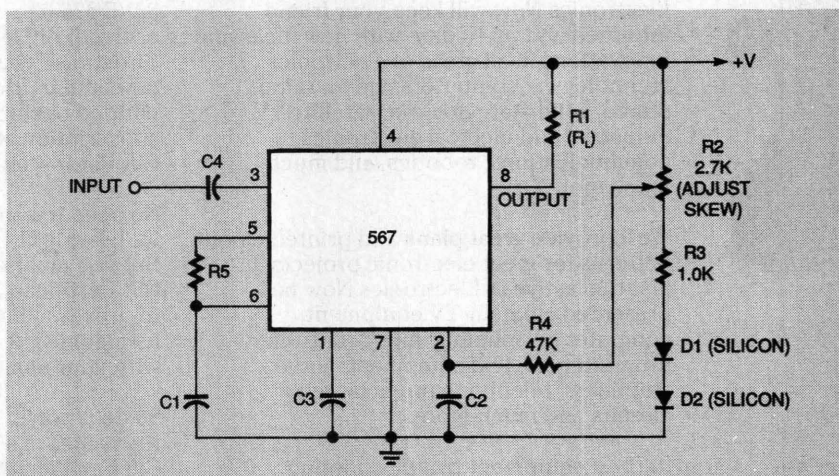


FIG. 13—TONE SWITCH WITH A TRIMMER potentiometer adjustment for skew.

R2. Capacitor C1 alternately charges through resistor R1, diode D1, and the left side of R2, and it discharges through resistor R1, diode D2, and the right side of R2 in each operating cycle. The operating frequency varies only slightly as the mark/space ratio is varied.

Figure 8 shows how the oscillator generates quadrature outputs. The squarewave outputs of pins 5 and 8 are out of

phase by 90°. In this circuit, input pin 3 is normally grounded. If it is biased above 2.8 volts, the square waveform at pin 8 shifts by 180°.

Figures 9 and 10 show how the oscillator circuit can be modified to allow timing resistor values to be increased to a maximum of about 500 kilohms. This permits the value of timing capacitor C1 to be pro-

(Continued on page 85)

TONE DECODER

continued from page 68

portionately reduced. In both circuits, a buffer stage is connected between the junction of resistor R1 and capacitor C1, and pin 6 of the 567.

In Fig. 9 this buffer is an emitter-follower transistor stage. Unfortunately, this stage causes a slight loss of waveform symmetry. By contrast, the circuit in Fig. 10 includes an operational amplifier follower as the buffer. It, however, causes no waveform symmetry loss.

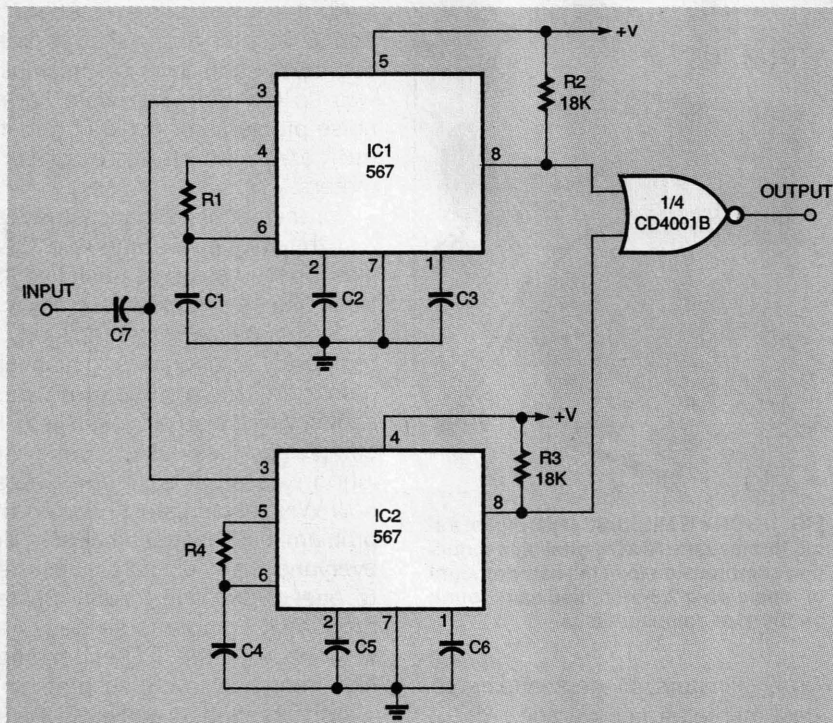


FIG. 14—DUAL-TONE DECODER with a single output.

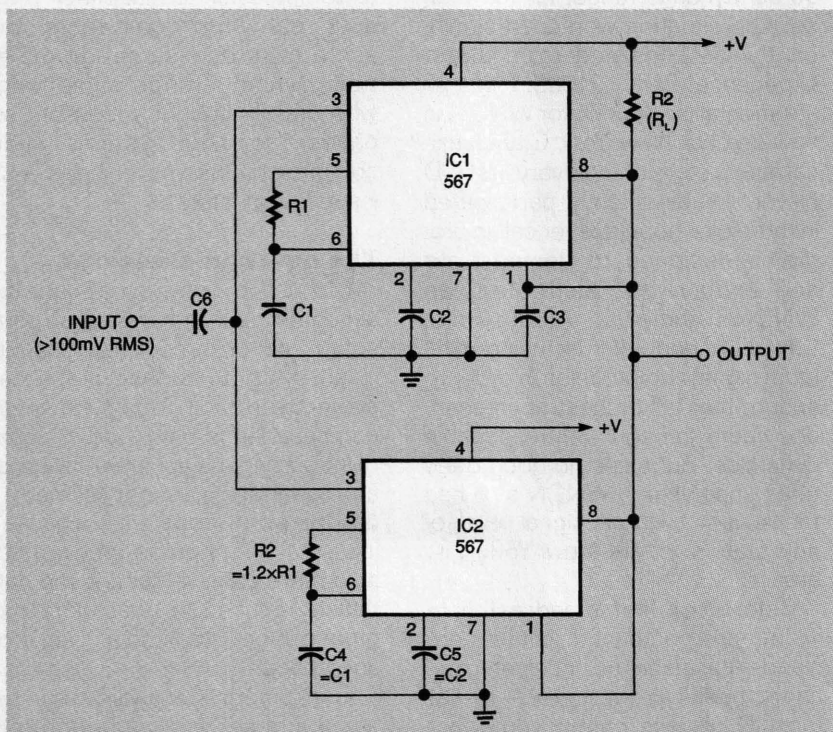


FIG. 15—DUAL-TONE SWITCH with 24 % bandwidth.

Five 567 outputs.

The 567 has five output terminals. Two of these (pins 5 and 6) give access to the oscillator output waveforms. A third (pin 8) functions as the IC's main output terminal, as previously stated. The remaining two outputs are available on pins 1 and 2 of the decoder.

Pin 2 gives access to the phase detector output terminal of the PLL, and it is internally biased at a quiescent value of 3.8 volts. When the 567 receives in-band input signals, this voltage varies as a linear function of frequency over the typical range of 0.95 to 1.05 times the oscillator's free-running frequency. It has a slope of about 20 millivolts per percent of frequency deviation.

Figure 11 illustrates the time relationship between the outputs of pin 2 and pin 8 when the 567 is organized as a tone switch. The relationships are shown at two bandwidths: 14% and 7%.

Pin 1 gives access to the output of the 567's quadrature phase detector. During tone lock, the average voltage at pin 1 is a function of the circuit's in-band input signal amplitude, as shown in transfer graph Fig. 12. Pin 8 at the collector of the output transistor turns on when the average voltage at pin 1 is pulled below its 3.8-volt threshold value.

Detection bandwidth

When the 567 is configured as a tone switch, its bandwidth (as a percentage of center frequency) has a maximum value of about 14%. That value is proportional to the value of in-band signal voltage in the 25 to 200 millivolt RMS range. However, it is independent of values in the 200 to 300 millivolt range, and is inversely proportional to the product of center frequency and capacitor C2. The actual bandwidth (BW) is:

$$BW = 1070 \sqrt{V_i / (f_o \times C2)}$$

in % of f_o and $V_i \leq 200$ millivolts RMS

Where V_i is in volts RMS and C2 is in microfarads

(Continued on page 88)

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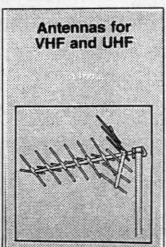
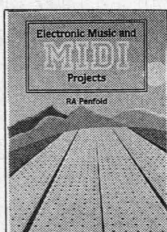
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TONE DECODER

continued from page 85

To select a C2 value by an educated trial and error process, start by selecting a value that is twice that of C1. Then either increase its value to reduce bandwidth, or reduce its value to increase bandwidth.

Detection band skew

Detection band skew is a measure of how well the band is centered about the center frequency. Skew is defined as:

$$(f_{\max} + f_{\min} - 2f_0)/2f$$

Where $f_{\max}$ and $f_{\min}$ are the frequencies corresponding to the edges of the detection band.

If a tone switch has a center frequency of 100 kHz and a bandwidth of 10 kHz, and its edge of band frequencies are symmetrically placed at 95 kHz and 105 kHz, its skew value is zero %. However, if its range of band values is highly nonsymmetrical at 100 kHz and 110 kHz, its skew value increases to 5%.

The skew value can be reduced to zero, if necessary, by introducing an external bias trim voltage at pin 2 of the IC with a trimmer potentiometer R2 and 47 kilohm resistor R4, as shown in Fig. 13. Moving the wiper up will lower the center frequency, and moving it down will raise it. Silicon diodes D1 and D2 are optional for temperature compensation.

Tone-switch design

Practical tone-switch circuits based on the typical connection diagram Fig. 3 are easy to design. Select the resistor R1 and capacitor C1 frequency control component values by referring to the nomograph, Fig. 6. Select the value of C2 on an empirical basis as described earlier. Start by making it twice the value of C1 and then adjusting its value (if necessary) to give the desired signal bandwidth. If band symmetry is critical in your application, add a skew adjustment stage, as shown in Fig. 13.

Finally, to complete the circuit design, give C3 a value double that of C2, and check the

circuit response. If C3 is too small, the output at pin 8 might pulsate during switching because of transients.

Multiple switching

Any desired number of 567 tone switches can be fed from a common input signal to make a multitone switching network of any desired size. Figures 14 and 15 are two practical two-stage switching networks.

The circuit in Fig. 14 functions as a dual-tone decoder. It has a single output that is activated in the presence of either of two input tones. Here, the two tone switches are fed from the same signal source, and their outputs are NORed by a CD4001B CMOS gate IC.

Figure 15 shows two 567 tone switches connected in parallel so that they act like a single tone switch with a bandwidth of 24%. In this circuit, the operating frequency of the IC2 tone switch is made 1.12 times lower than that of the IC1 tone switch. As a result, their switching bandwidths overlap. Ω

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Compressors, expanders, and companders

ROBERT F. SCOTT, SEMICONDUCTOR EDITOR

SINCE THE EARLY DAYS OF HIGH-FIDELITY audio, engineers have worked to improve the realism and signal-to-noise ratio of both recorded and broadcast music. Recording engineers, however, often limit or compress the dynamic range, and broadcasters limit or compress the signal amplitude, of that music. That is done to prevent overloading a tape or overcutting a record, and to prevent overmodulation. However, those same efforts cause the full dynamic range of the original music to be lost to you—if your playback system does not include a *dynamic volume expander*.

The Signetics NE570 compandor

You can build a professional-quality expander, compressor, or compandor (a combination compressor and expander circuit) for your hi-fi system by using circuits designed around the Signetics NE570 IC compandor. The pin-out of the IC is shown in Fig. 1. As a compressor, the device provides a 2:1 compression ratio—for example, a 100-dB dynamic range of +20 dB to -80 dB is com-

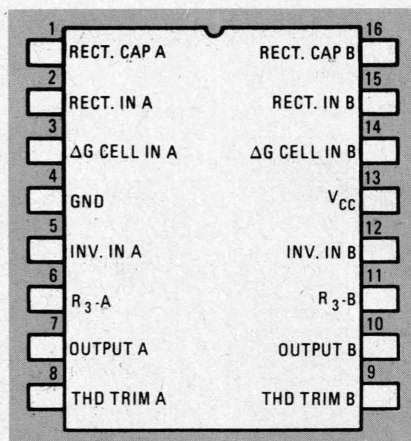


FIG. 1

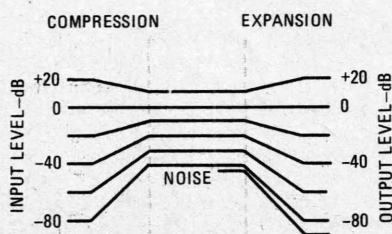


FIG. 2

pressed into a 50-dB (+10 to -40 dB) range as shown in Fig. 2. As an expander, it has a 1:2 expansion ratio, taking the +10 to -40 dB compressed signal and restoring its original full dynamic range.

A compandor can be used for noise reduction. In that application, the signal is compressed before noise can be introduced, and expanded afterwards. Figure 2 shows how that method of companding (compressing and then expanding) can improve the signal-to-noise ratio by about 45 dB.

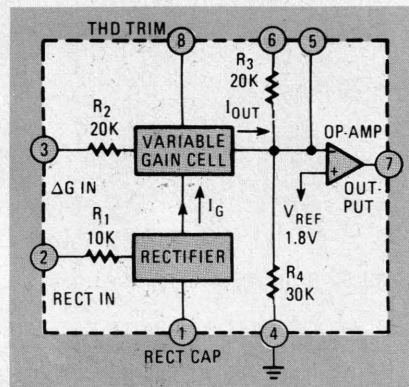


FIG. 3

A block diagram of one half of the NE570 is shown in Fig. 3. Each half of the IC consists of a full-wave rectifier, a variable-gain cell (ΔG), an op-amp, and a biasing system. The full-wave rectifier and an external capacitor (tied to the RECT CAP terminal) detect the average value of the input signal. The rectifier output current (I_G) controls the gain of the variable-gain cell. Therefore, the gain of that section of the circuit is proportional to the average value of the input-signal voltage. The ΔG output current, I_{OUT} , is fed to the inverting input of an on-chip op-amp that is biased at V_{REF} . That reference voltage is 1.8 volts and is provided by a very stable internal low-noise source. (That internal precision voltage-source also biases the THD TRIM circuit used for temperature compensation.)

The speed with which the circuit gain can follow changes in the amplitude of the input signal depends on the value of the external capacitor (the one attached to the RECT CAP terminal). A small capacitor will provide fast attack and fast decay times, but may not provide enough low-frequency filtering. In that case, residual

low-frequency signal components will appear on I_G and will modulate the signal passing through the variable-gain stage. That results in third-harmonic distortion, so there must be a compromise between fast response and distortion.

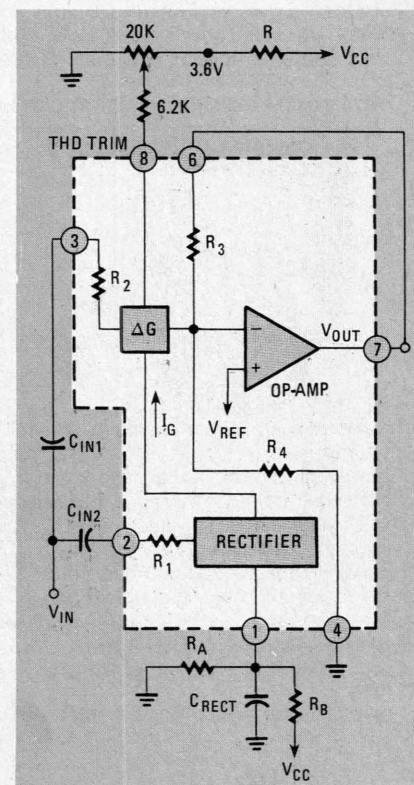


FIG. 4

The expander's output is determined by the DC gain provided by the op-amp. The output is related to the internal reference voltage and also to biasing resistors R_3 and R_4 as expressed by the equation: $V_{DC OUT} = (1 + R_3/R_4) V_{REF}$. When V_{CC} (the supply voltage) is higher than 6 volts, R_4 should be shunted with an external resistor to bias the output up to $\frac{1}{2} V_{CC}$.

Resistor R_3 is brought out from the op-amp summing node and is used when you want expander or compressor gain to be set solely by on-chip components. You can adjust that gain to your needs by placing external resistors in series with R_3 . You can also connect an external resistor across R_4 to change the bias to any value desired.

The basic expander

Figure 4 shows the circuit of a basic expander. Input signal V_{IN} is applied to the rectifier and ΔG stage inputs in parallel. The expander can handle a signal input up to 3 volts peak. Rectifier input current can be as high as 300 μA , while the input to the ΔG stage should be limited to 140 μA . If the compressor will see input signals greater than +2.8-volts

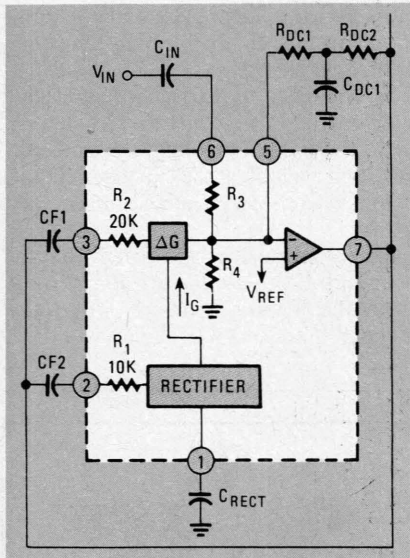


FIG. 5

peak, use suitable resistors in series with R_1 and R_2 to limit currents to the specified values.

Voltage offsets in the ΔG stage can cause distortion; primarily even harmonics. The THD TRIM pin permits a compensating external voltage to be applied to neutralize the effect of the offset voltages. A voltage divider composed of a 20K pot and series resistor R is connected between V_{CC} and ground as shown in Fig. 4. A 6.2K resistor is connected to the THD TRIM pin. The value of resistor R is selected to develop 3.6 volts at the high end of the pot.

In Fig. 4 coupling capacitors are shown in series with both the rectifier and ΔG stage inputs. However, R_1 and R_2 can be tied together and connected to the signal input through a single coupling capacitor. In that case, though, tracking at low input-signal levels will be degraded.

The comparator transfer-tracking tends to be a linear 2:1 ratio down to a very low input-signal level. Then, tracking may deviate in either direction from the normal 2:1. Either resistor R_A or R_B (but not both) may be needed to adjust transfer linearity. To correct low-level tracking error, select a suitable value for R_A ranging from around 1 megohm to 100K or for R_B between 250K and 5 megohms.

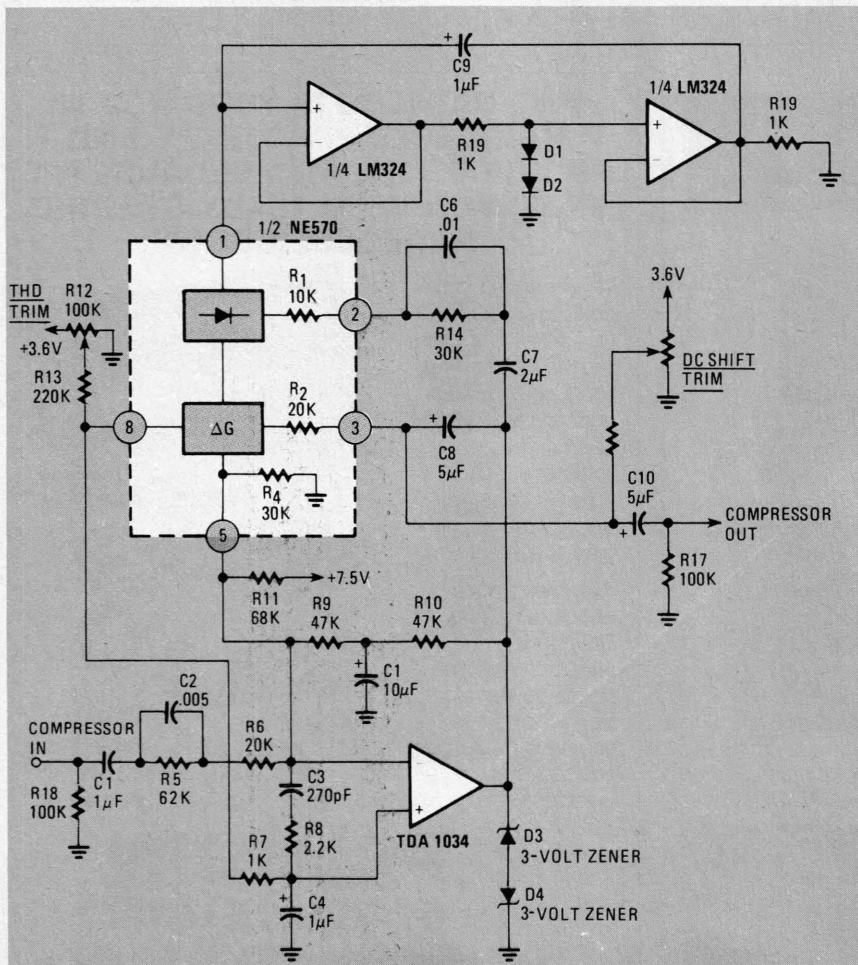


FIG. 6

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The basic compressor

Figure 5 shows that the dynamic compressor is essentially an expander inserted in the feedback loop of an op-amp. The inputs of the ΔG stage and the rectifier are tied to the op-amp output. The variable-gain stage is set to provide AC feedback only; DC feedback is provided by an external low-pass network composed of R_{DC1} , R_{DC2} , and C_{DC1} . The sum of the values of the two feedback resistors determines the bias at the op-amp's output. The output voltage $V_{DC OUT}$ can be written:

$$V_{DC OUT} = 1 + \frac{R_{DC1} + R_{DC2}}{R_4} V_{REF}$$

$$= \left(1 + \frac{R_{DC TOT}}{30K}\right) 1.8V$$

When internal bias resistors R_3 and R_4 are used alone, the expander output will be:

$$V_{DC OUT} = 1 + \frac{R_3}{R_4} V_{REF}$$

$$= \left(1 + \frac{20K}{30K}\right) 1.8V = 3.0V$$

You can shunt a suitable resistor across R_4 to raise the output bias to the desired level; and you can connect a resistor in series with R_3 to increase op-amp output gain.

For the widest possible dynamic range, the compressor's output-level should be as high as possible. Therefore, the input

to the rectifier should be as high as possible without exceeding the $+300 \mu A$ peak-current limit. If the average input-signal level is low, a higher output can be obtained by using a shunt resistor to reduce the effective value of R_3 or by using an external series resistor to increase the effective value of R_2 . Note well that a reduction in the effective value of R_3 reduces the circuit's input impedance.

A high-fidelity compressor

Figure 6 shows a circuit for a hi-fi dynamic compressor that would make an ideal accessory for your tape-recording setup. It features high gain and wide bandwidth. Its external rectifier-capacitor (C_9) is not grounded. Instead, it is connected to the output of an op-amp network (IC1-a and IC1-b) to shorten the compressor attack-time at low signal-levels. (The attack time of the basic circuits in Figs. 4 and 5 is relatively long.) That external op-amp is used to provide improved high-frequency gain.

Diode D3 and D4 clamp the compressor output to a 7-volt peak-to-peak swing. That is necessary at times when the compressor is operating near maximum gain—as with a small signal input—and is suddenly hit with a high-level signal. Normally, the output would swing from V_{CC} to ground and would overload the circuit the compressor was feeding—a tape recorder for example.

The attack time and the time it takes for the compressor to recover from an overload depend on the value of C_9 . A value of about $1 \mu F$ is a good compromise.

Breathing

Even some of the best broadcast-quality compressors have been said to have a problem with *breathing*. That term refers to slow cyclic variations in background level that can be heard as the compressor changes gain. Breathing is minimized in this circuit by high-frequency pre-emphasis networks C2-R5 and C8-R14. Naturally, the expander should have a de-emphasis network to complement the compressor's pre-emphasis network. We'll take a look at the expander circuit, before we go on to other things, next month.

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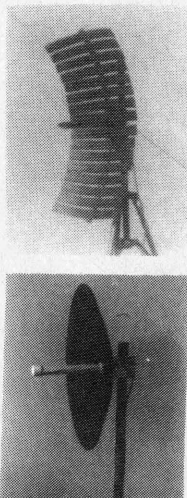
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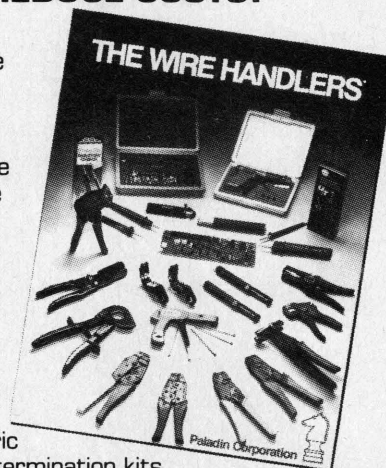
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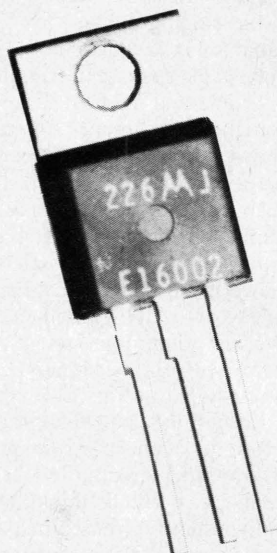


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How to Design Analog Circuits Audio Power Amplifiers



MANNIE HOROWITZ

Here's a look at some practical audio power-amplifier circuits. Circuits using both bipolar and FET devices will be covered.

ALTHOUGH IN THE PAST MANY PIECES OF audio equipment used transformers to couple the driver stage to the power transistors, and those transistors to the loudspeaker, output transformers are currently used only in equipment providing very low output power. You are likely to find an output transformer in a portable radio, but in little else. As for sophisticated equipment, economy may dictate that a driver transformer be used, but output transformers are usually avoided because they may severely limit the fidelity of the signal delivered to the loudspeaker. Instead, most modern audio equipment uses one of a variety of types of transformerless circuits to drive the power-amplifier stages.

Transformerless amplifiers have in the past mainly used bipolar power-transistors. The present trend, however, is to use power VFET's and MOSFET's. One reason for that is the absence of problems such as thermal runaway and second breakdown inherent in bipolar transistors. Another important reason is that FET characteristics are more linear than those of their bipolar counterparts. Consequently, when amplifiers using FET's as output devices are compared with those

using bipolar transistors, the distortion is lower in the FET circuits. As a result, you need less feedback to reduce distortion to near ideal levels with FET amplifiers than you would in bipolar amplifiers. And, because less feedback is required in FET amplifiers, instability problems due to feedback are less.

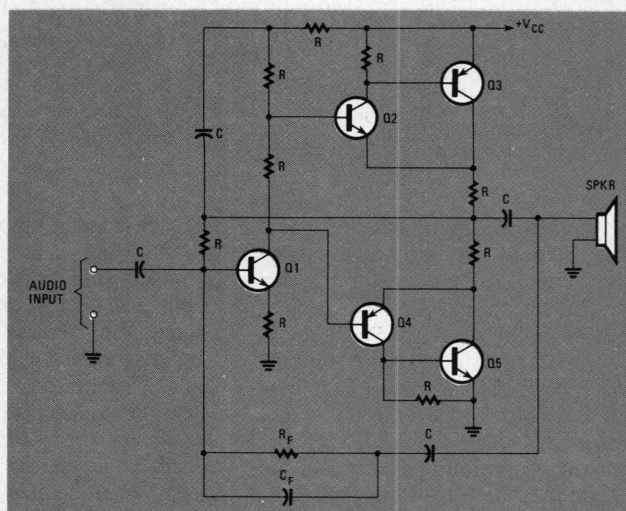
Driver transformer circuits

A circuit using a driver transformer is shown in Fig. 1. The input signal is fed to the base of Q1 and amplified. The amplified output appears across the primary winding (winding 1) of the driver transformer, T1. The signal from that winding is induced into the two secondary windings and applied from there to output transistors Q2 and Q3. Note that in Fig. 1 there is a dot shown at one end of each secondary. Those dots indicate which ends of the various windings are in phase. While a signal is applied to the base of output transistor Q3 from the end of winding 3 with the dot, a signal of the opposite phase is applied from winding 2 to the base of Q2 from the terminal without the dot—in other words, the same signal is applied out of phase to the two output transistors. If the transistors were biased

so that they did not conduct when idling, each transistor would conduct only when a signal was present—in this case only during alternate halves of the cycle. The outputs from Q2 and Q3 will then combine across the loudspeaker load to reproduce the original signal.

Transistors are not biased for zero idling current. There is always some current flowing so that the output devices operate in Class-AB. Bias current for Q2 flows through R3 and through winding 2 of the transformer to the base. Although some of the current from R3 is diverted through R4, there is sufficient current left for the base of Q2 to keep it turned on while idling. A similar arrangement involving R5 and R6 keeps Q3 turned on.

Resistors R7 and R8 in the emitter circuits of Q2 and Q3 respectively are not used exclusively in circuits with driver transformers. They are frequently found in completely transformerless circuits. Those emitter resistors increase the voltage gain of the driver stage while significantly reducing the voltage gain of the output devices. To minimize that loss of gain, the values of the resistors are kept small, and the circuit is designed so that between 0.5 and 1 volt is across each of



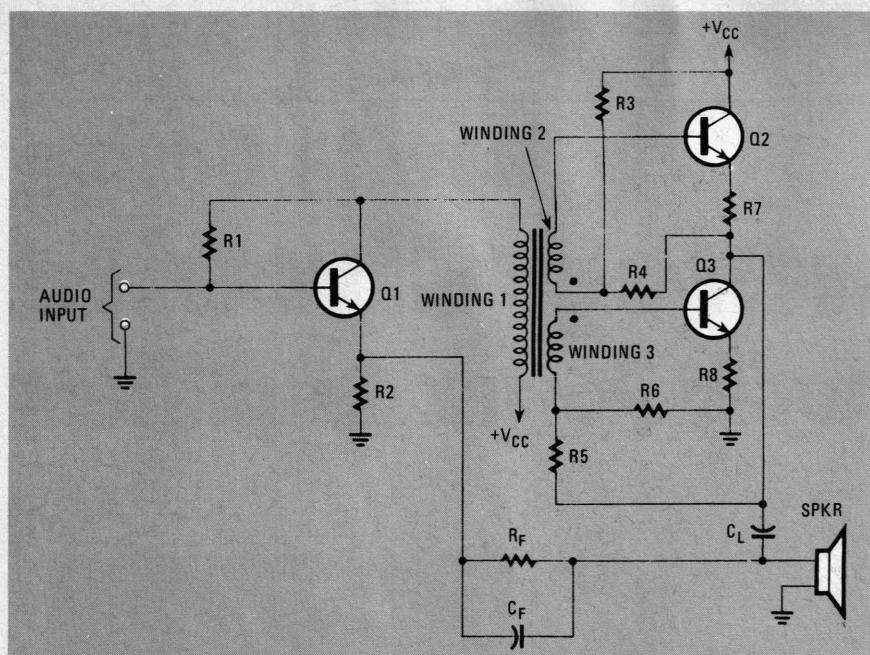


FIG. 1—POWER AMPLIFIER with driver transformer. The outputs from Q2 and Q3 recombine across the speaker to reproduce the input signal

the resistors when the transistors idle.

As is true with just about every other transistor circuit using a resistor in series with the emitter, resistors R7 and R8 help to stabilize both AC gain and DC bias. Those resistors also serve as output-transistor protection devices. That protection is important because an output transistor may break down if a short develops across the loudspeaker. But the protection that those resistors provide is somewhat limited; more complex feedback circuits do a better job.

In Class-AB push-pull amplifiers, during different portions of the cycle either one transistor or the other conducts more heavily. During one half cycle, Q2 may conduct heavily and Q3 may not conduct at all, while in the other half the situation may be reversed. In each cycle, however, both transistors must change from a conducting to a non-conducting state and vice versa. Resistors R7 and R8 help to make that transition smooth, keeping crossover distortion to a minimum. To really improve the smoothness of the transitions, diodes can be substituted for the emitter resistors.

Driver transistor Q1 supplies the bulk of the voltage gain for the circuit while providing sufficient power to drive output transistors Q2 and Q3 through the transformer. The turns ratio of the transformer is selected for minimum distortion across the output load, and is found by trial and error. Typically, however, the turns ratio is usually about 1.7:1. If transformers are not readily available for substitution into the circuit, you will have to live with what you do have but add a feedback circuit to reduce distortion to reasonable levels.

Let's now see how feedback can be used to reduce distortion. The signal is

fed back from the output to Q1 through R_F and C_F . If the phase is proper, the voltage gain of the circuit is reduced when those components are connected as shown. (Should gain increase or should the circuit oscillate, improper phasing is usually at fault. To correct that situation, just reverse the connections to the primary of the driver transformer.) The network adds what is referred to as negative feedback. When the gain is reduced so is the distortion. If gain is reduced too much, however, the circuit may oscillate. You can determine the amount of usable feedback by trial and error—by varying both R_F and C_F .

A circuit may become marginally unstable even when negative feedback is added. That is because feedback may be negative within a specific frequency range (the range in which the quantity of feedback is being measured) but become positive outside of that range. A square-wave generator and an oscilloscope can be used to check the stability of an amplifier with feedback. Start by feeding a 10-kHz squarewave to the input of the

amplifier. Note the waveform across the amplifier's output—it should be reasonably square. The three displays that you are most likely to see are shown in Fig. 2. In Fig. 2-a, the ringing on the top and bottom of the squarewave tends to rise with time while in Fig. 2-b it decreases. In Fig. 2-c, there is no ringing, but the leading edge of the squarewave is rounded.

When the output is as shown in Fig. 2-a, the circuit has a tendency to oscillate. That is indicated by the rising amplitude of the ringing signal. Even though the signal in Fig. 2-b also shows ringing, it is more stable because the ringing decreases with time and tends to disappear. To go from the state shown in Fig. 2-a to the one shown in Fig. 2-b usually involves simply increasing the value of C_F . If, however, C_F is made too large, ringing may be eliminated but the leading edge of the squarewave will become rounded as shown in Fig. 2-c. If that happens, there may be a loss of high frequency response. The best compromise to adjust C_F so that the waveform is somewhere between those shown in Figs. 2-b and 2-c.

Do not disregard the information presented here concerning the proper design of transformer-coupled circuits with feedback. You may think that it does not apply when no transformer is used, but that is not true. The information presented here applies to all types of power amplifiers. As for feedback, the details and characteristics will be covered in a later article in this series.

Amplifiers using a complementary circuit

For best results from a push-pull circuit, the two halves of the output circuit must be identical. That is not the case in the circuit shown in Fig. 1. There, the output from Q2 is taken from its emitter while the output from Q3 is taken from its collector. Consider, on the other hand, the circuit shown in Fig. 3. In that transformerless circuit, transistors Q2 and Q3 (NPN) and transistors Q4 and Q5 (PNP) form two darlington pairs.

The loudspeaker load is fed by one Darlington pair during the first half of the cycle, and the other one during the second so that the output signal is perfectly

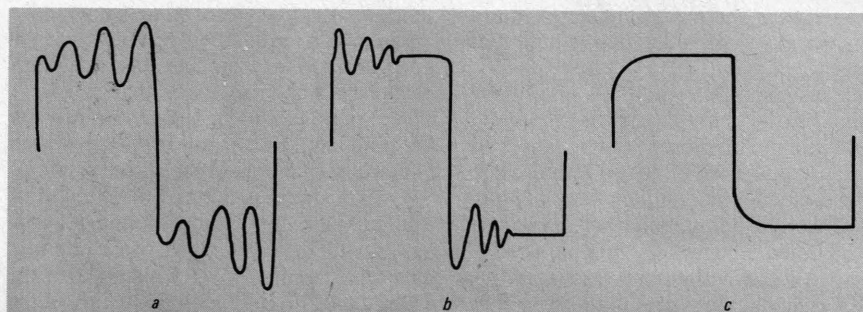


FIG. 2—IF A SQUAREWAVE is applied to the input of an amplifier, the waveforms shown here may be observed at the output. The waveforms in a and b indicate oscillation (ringing); the one in c indicates loss of high-frequency response. All of those conditions can be changed by changing the value of C_F .

balanced. The input signal is applied to the voltage amplifier, Q1. Its amplified output is applied, in phase, to both Q2 and Q4. That phase relationship is necessary for Q2 to conduct on positive halves of the cycle and Q4 on negative halves. This same phase relationship exists at the outputs from Q3 and Q5. Those outputs are recombined across the loudspeaker to reproduce the full cycle.

Some idling current must flow in the output transistors, just as it did in the circuit in Fig. 1. The amount of idling current is established by the voltage developed across R3. That voltage, when applied to the bases of Q2 and Q4, produces a current through the base-emitter circuits of those transistors. Emitter currents from Q2 and Q4 set the base currents and hence idling emitter currents of Q3 and Q5. In most instances, R3 is replaced by forward-biased diodes, or a series resistor-diode combination, to stabilize idling current despite temperature variations.

But how is a voltage developed across R3? (That is one of the load resistors in the collector circuit of Q1.) The others, wired in series with R3, are R4 and R5. When collector current flows through Q1, the voltage required to forward bias Q2 and Q4 is developed across R3. Transistor Q1 is biased through resistor R1, which is connected to the junction of R8 and R9. When idling, the voltage at that junction is ideally $\frac{1}{2}$ of $+V_{CC}$. Resistor R1 is connected to that point to help stabilize the bias of Q1 against temperature variations.

In order to minimize distortion, a considerable amount of negative feedback must be used around the circuit. If a lot of feedback is applied, however, the gain will drop to low levels. To compensate for that, the forward gain of Q1 must be made very high. Capacitor C_B helps the circuit meet that gain requirement. Signal is fed back through C_B from the output to the junction of R4 and R5. That is known as a "bootstrapping" circuit. That bootstrap circuit makes R4 appear to be much larger than it actually is. And, as R4 is part of the collector load-resistance, the forward gain of Q1 is very high because it is approximately equal to the ratio of the resistance in its collector to the resistance in its emitter.

Capacitor C_B also serves a more important purpose. When the signal is large, the emitter of Q2 is at $+V_{CC}$ volts. When that happens, no current can now flow through its base-emitter junction because the emitter is more positive than the base and Q2 does not conduct. Peaks in the signals are consequently cut-off causing distortion. Let's see how including C_B in the circuit corrects that situation. That capacitor is charged to about $\frac{1}{4}$ of $+V_{CC}$ when the circuit is idling. When a peak is present in the signal, not only is the emitter of Q2 at $+V_{CC}$, but since the bottom of C_B is effectively at the same potential

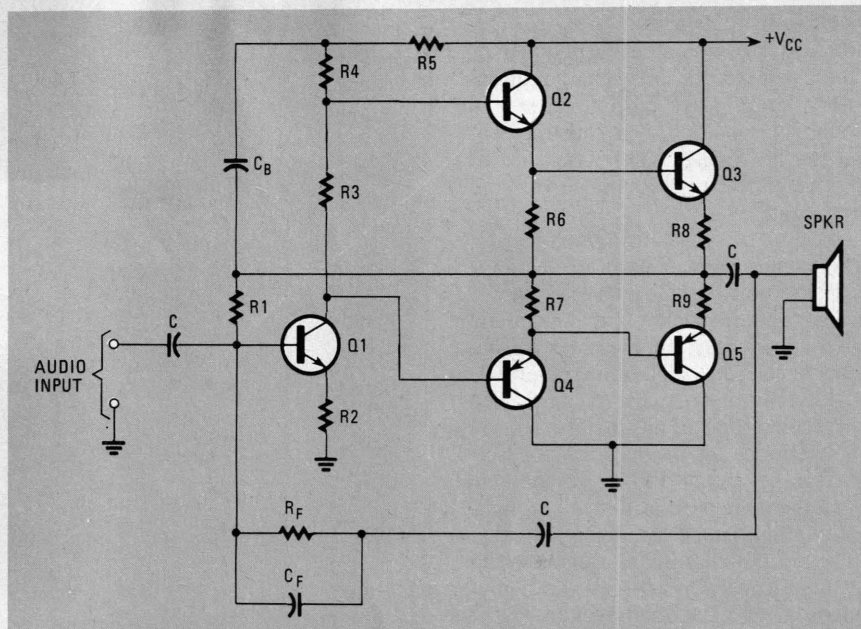


FIG. 3—DARLINGTON PAIRS are used in the output circuit of this audio power-amplifier.

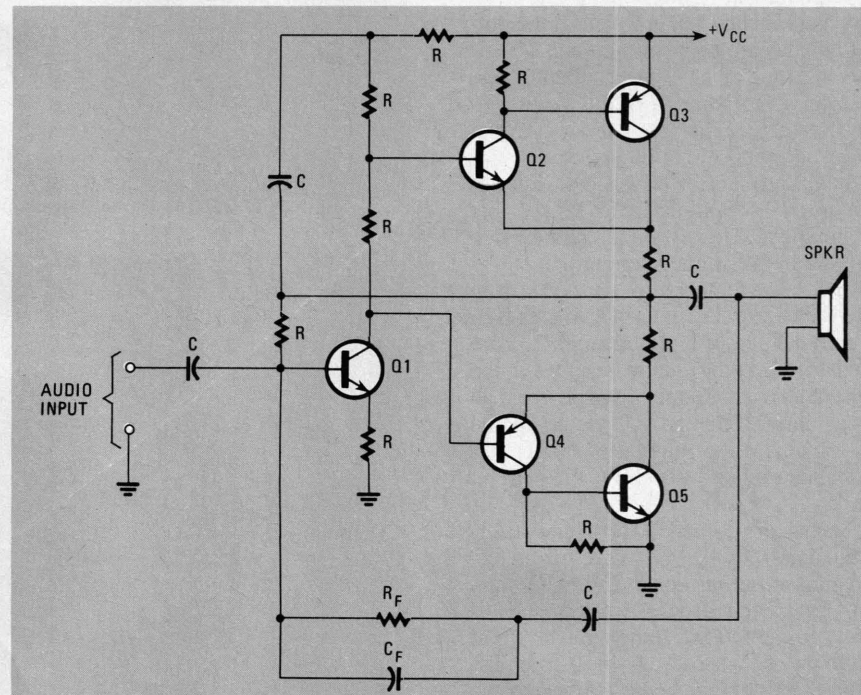


FIG. 4—COMPLEMENTARY PAIRS are used in the output circuit of this amplifier. The signals from them combine across the speaker to reproduce the input signal.

as the emitter, that terminal of the capacitor is also at $+V_{CC}$. Because the capacitor is charged to about $+V_{CC}/4$, the top terminal of C_B is at $+V_{CC} + V_{CC}/4$. That voltage is applied to R4 to make the base of Q2 positive with respect to its emitter, turning Q2 on. Being turned on, peak positive pulses can now pass through Q2, and the balance of the circuit, to the loudspeaker.

When the circuit is idling, C_B does not affect the performance of the amplifier. Resistors R4 and R5 are chosen so that base current in Q2 and Q4 is proper for the desired idling current to flow through the output transistors. The values of R4 and R5 are usually identical. As before, R_F and C_F form the negative feedback cir-

cuit. The method used to find the values for those components are identical to the one previously discussed.

Complementary circuits can be used in place of the Darlington pairs in the power-output circuit. The complementary pair was described in the article on coupled circuits. A circuit using complementary pairs is shown in Fig. 4. Here, Q1 performs the same function as it did in the circuit shown in Fig. 3. Transistors Q2 and Q3 form one complementary pair; transistors Q4 and Q5 form a second.

One of the big drawbacks of the two transformerless circuits discussed thus far is the presence of a capacitor between the output circuit and the loudspeaker. That capacitor must have a high value if it is to

pass the low frequencies. Since the capacitor gets charged through the output transistors, and since the initial charge current is very large, more current may flow through Q3 and/or Q5 at that moment than can be handled safely. Because of that, one or both of those transistors may break down.

A second drawback using that capacitor is that it is almost always an electrolytic because of the high values required. An electrolytic capacitor is not linear, and consequently just the presence of that capacitor can add to distortion somewhat.

The circuit shown in Fig. 5 can be used to overcome some of those drawbacks by simply eliminating the need for a capacitor. Arrangements similar to the one shown there are used in some very high-quality amplifiers.

The big problem in amplifiers that do not use a capacitor between the output transistors and the loudspeaker is that there is no way of keeping DC from flowing through the speaker. The circuit in Fig. 5 eliminates that problem. If the output devices are connected to equal positive and negative voltage supplies, the voltage at the junction of the output devices is zero. That assumes that equal idling current flows through the two complementary pairs of transistors. Current can usually be adjusted to satisfy that requirement. However that relationship will hold only at one temperature; it will not when the temperature rises or falls in the preceding DC-coupled stages. To overcome that, differential amplifiers are used to drive the output stages—if the current changes in one of the devices, an equal current change will occur in the second device, keeping the overall circuit in balance. Let's see how that circuit works.

Transistors Q1 and Q2 form one differential amplifier. They drive a second differential amplifier consisting of Q3 and Q4. The output from Q3 is applied directly to the Q6/Q7 complementary pair while the signal from Q4 must first pass through Q5 before being applied to the Q8/Q9 complementary pair. Transistor Q5 is required because it shifts the phase of the signal from Q4 so that the signal fed to Q6 is in phase with that at the input of Q8. Resistors in the base and emitter circuits of Q5 are adjusted so that the current from Q5 is equal to the current from Q3. No bootstrap capacitor is required in that circuit as the proper current levels are always present at Q6 and Q8, through Q3 and Q5 respectively.

Potentiometer R1 is adjusted so that there is 0 volt at the junction of Q7 and Q8, and across the loudspeaker. Transistor Q10 is in a constant current source circuit, required for proper operation of the differential amplifier.

The circuit shown in Fig. 6 is similar to the one in Fig. 5. The op-amp, as discussed in a previous article, is actually a combination of differential amplifiers.

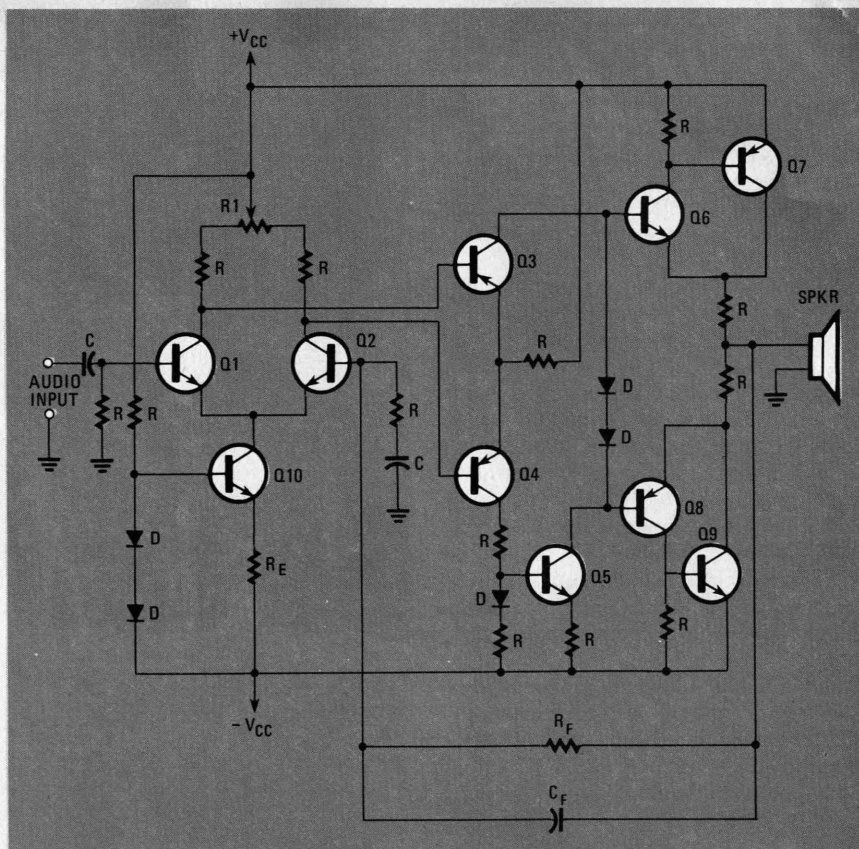


FIG. 5—THE HEART OF THIS high-quality circuit is a pair of differential amplifiers.

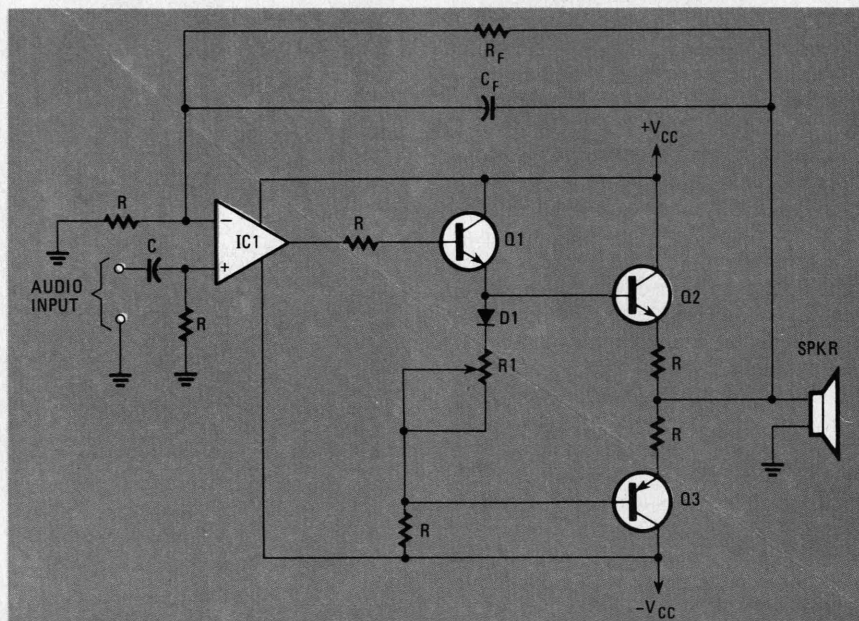


FIG. 6—SINCE OP-AMPS are simply combinations of differential amplifiers, they can be used in this variation of the circuit shown in Fig. 5.

As such, its DC-output level is extremely stable despite temperature changes. Because that stable voltage is coupled to the output devices, a loudspeaker can be connected directly to those output transistors without an intervening capacitor.

Note two items peculiar to this circuit. Instead of using Darlington or complementary pairs in the output, a single output transistor is used in each leg of the push-pull circuit. Second, the voltage de-

veloped across D1 and R1 is used to establish the bias for Q2 and Q3. The desirable idling current for the output transistors is set by adjusting R1 because that potentiometer varies the voltage applied to the base circuits. Diode D1 helps keep that voltage, and hence the idling current, constant despite variations in temperature.

Next month we'll continue our discussion of power amplifiers.

R-E

ALL ABOUT

Part 2 IN THE FIRST ARTICLE of this series, we presented some of the fundamentals of active receiving antennas. That type of antenna has several advantages over wire antennas, especially at very-low and low frequencies (VLF and LF). First, active antennas have a short physical length. The active antenna systems that we will discuss here are used with a one-meter long whip. That helps reduce the sensitivity to local noise from sources such as power lines. Because of the active antenna's high input-impedance and low output-impedance, it is more efficient than a simple wire antenna in converting a received signal at the antenna to a corresponding voltage level at the receiver's antenna terminals.

In general, the properties that we want our active receiving antenna to have are: high input-impedance, low input-capacitance, low output-impedance, and minimum distortion/high linearity.

Another objective is to keep the circuit as simple as possible. A single-stage JFET amplifier has the best combination of properties for active antenna preamplifier applications—and it allows the circuit to be kept relatively simple. (This is not to suggest that there might not be better, more complex circuits, using several semiconductors or IC's.)

Wide-band amplifier circuit

The JFET that we have chosen to use is the Siliconix J-310 (or U-310 in metal can). That JFET is often used as a grounded-gate transmission-line amplifier for TV and FM reception (at a 75-ohm input/output level). The J-310 will usually handle short-duration static surges up to 100 volts or so without damage, so a

VLF Active Antennas

An active receiving antenna can dramatically improve your receiver's performance, especially at very low frequencies. Here we will discuss some practical circuits for both wideband and narrowband operation.

single low-capacitance neon bulb can provide input static-charge protection. That is of value since semiconductor diodes usually have a much higher junction-capacitance when used as protection devices and, if used, would increase the input capacitance of the preamplifier.

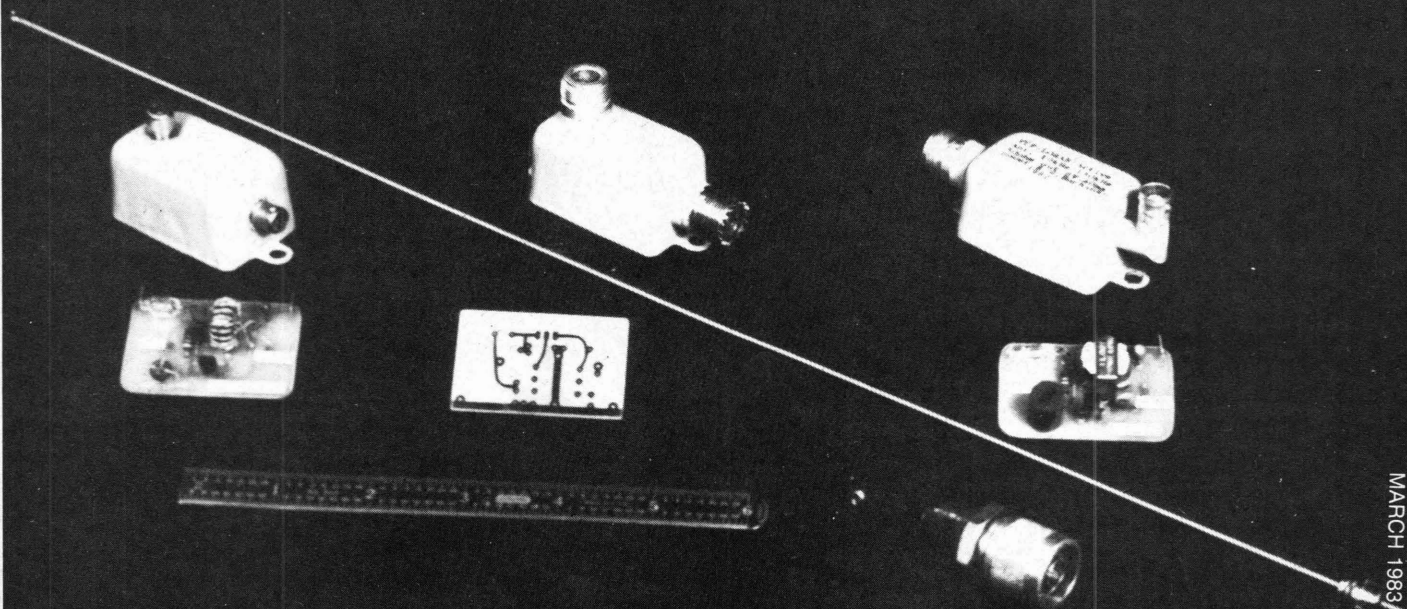
In our application as an active VLF-HF preamplifier, the J-310 is used in a common-source common-drain configuration with inductive feedback (that improves the linearity and lowers the output impedance). Figure 1 shows our wideband circuit for the range of 10 kHz to 30 MHz. Note that the feedback from drain to source is large because of the low resistance of the transformer and its 1:1 turns ratio. (We will discuss how to wind that transformer in Part 3 of this series; that part will contain actual construction details.) For the circuit to operate properly, the transformer's output should be opposite in phase to its input (with respect to ground).

The amplifier circuit is intended to be used with a 1-meter vertical whip. The antenna and its mount capacitances serve as part of an input filter. The input capacitance of the JFET is quite low (about 7 pF). The 2.2- μ H inductor at the gate of the JFET serves as a lowpass filter or trap, resonating with the junction and circuit (including antenna) capacitances at a frequency near 30 MHz. That input filter aids in reducing FM-VHF interference over a range of 50 to 500 MHz where the 1-meter whip acts like a resonant antenna.

Receiver coupler

The receiver coupler both provides power to the preamplifier and extracts the signal from the coaxial transmission line (from the preamp). A wideband receiver

R.W. BURHANS



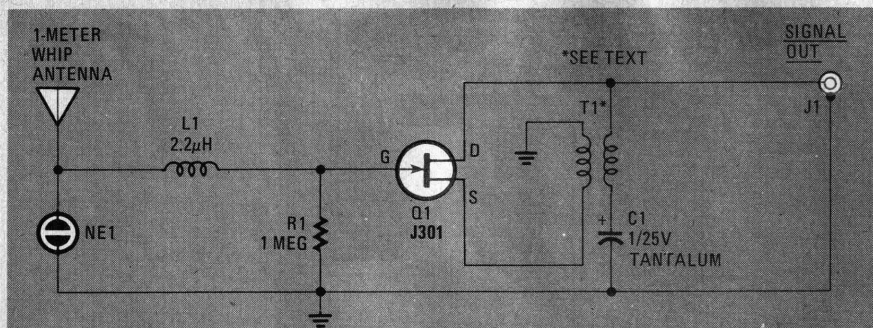


FIG. 1—THE WIDEBAND AMPLIFIER. The transformer should be connected so that the polarity of the output is opposite in phase to that of its input.

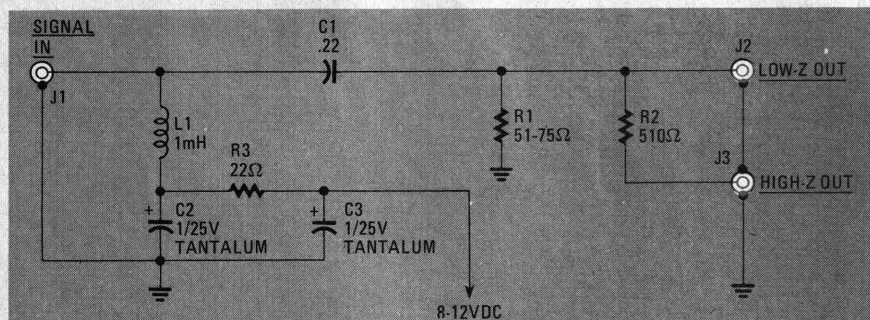


FIG. 2—THE RECEIVER COUPLER both provides power to, and extracts signals from, the amplifier, as well as acting as a highpass filter

coupler is shown in Fig. 2. Capacitor C1 and inductor L1 form a highpass L-section filter (with about a 10-kHz 3-dB rolloff). Resistor R1 is used to ensure that the preamplifier output sees a low-impedance load no matter what sort of receiver is connected. Resistor R2 is used for matching to a receiver with a higher input impedance. That resistor would cause a signal loss of 6 dB if the input impedance to the receiver were 500 ohms.

The coupler circuit provides DC power to the preamp through the coaxial cable. Power sources less than about +8 volts will reduce the dynamic range and linearity of the amplifier. The power dissipation of the JFET using a +8-volt supply will be about 200 mW. The rating of the J310 at 25°C is about 360 mW maximum. In practice, we have not burned one up even when operated with a +12 volt supply for an extended length of time.

The active antenna preamp is like a Class-A amplifier (where the output has low distortion, but the power furnished by the DC power supply is much greater than the power dissipated in the load). However, some distortion does ultimately appear in the output at high input-signal levels. That is due to the fact that a JFET biased in that way cannot be made perfectly linear over a wide dynamic swing of the output voltage. Other modes of operating the JFET with different biasing have been tried, but they have not resulted in any significantly better performance. So, in a sense, the circuits of Figs. 1 and 2 are of the "simpler is better" type.

Intermodulation distortion

A wideband active antenna covering

from 10 kHz to 30 MHz has poor performance with regard to IMD (InterModulation Distortion) because little input filtering is provided. Interference will be noted especially if the observer is close to strong AM broadcast-band transmitters. The standard method for evaluating the intermodulation response

of a receiver is to measure the 2nd and 3rd order intercepts.

Figure 3 shows a plot of the output power of the two fundamental signals (f_1 , f_2) versus the output power of the second order and third order distortion products. (We discussed intermodulation distortion products in the first part of this series, which appeared in the February issue of **Radio-Electronics**). Those are shown as a function of the power of a two-tone input signal.

One thing we should mention first is that when the input signals are too large, the amplifier output will not follow the input linearly. That is called *gain compression* and can be seen in Fig. 3.

If the linear portions of the curves are extended, they will eventually cross each other. That is shown in Fig. 3, where the curves are extended by dotted lines and cross at an output level that cannot be reached by the amplifier. The point where they cross is called the *amplifier intercept*. The input and output coordinates where they cross give you the input and the output intercepts.

In general, the higher the intercept point is on the graph, the better the amplifier's capability. Those measurements are best made with a sensitive spectrum analyzer, but an approximate idea can be obtained by using a receiver and recording the S-meter readings with appropriate signal-generator sources. The relatively low number of only +10 dBm for the 3rd order intercept indicates that the active antenna should be used

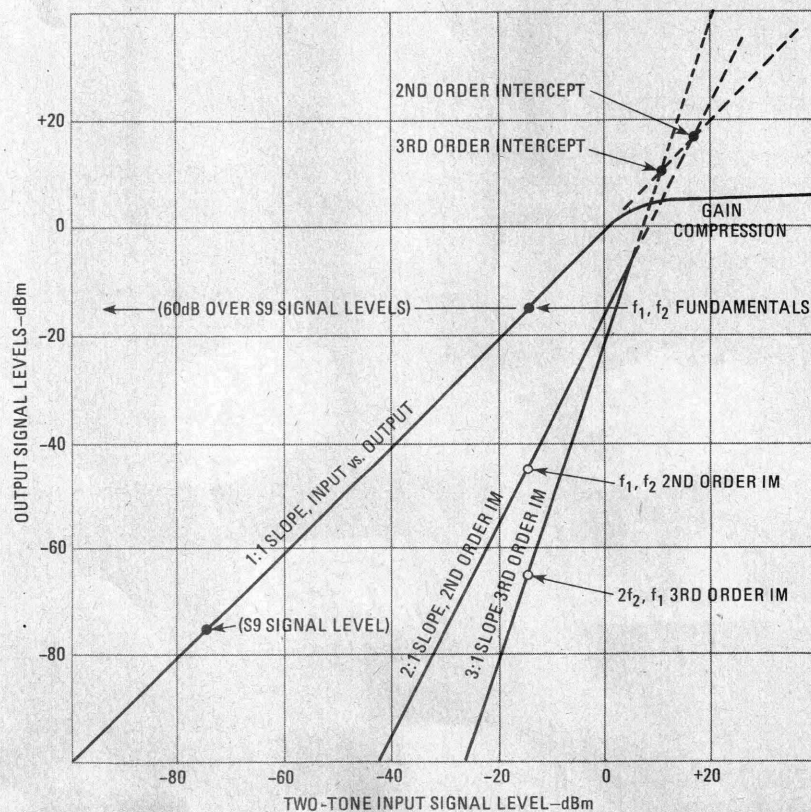


FIG. 3—THE HIGHER THE INTERCEPT POINTS, the better the amplifier's intermodulation rejection.

over a wide frequency range only where the local interference level is not severe. The antenna, of course, might be used in a high-signal area but the observer has to exercise some caution in making sure that the IM signals are not obscuring some desired signals on the same frequency.

For the wideband case of 10 kHz to 30 MHz, those intermodulation-distortion measurements suggest that only a short antenna of perhaps 1 meter or even less will provide the least amount of spurious responses—increasing the antenna length will only tend to increase the distortion level. Longer antennas should be used only when the active preamplifier is provided with some form of input and/or output filtering to reduce the out-of-band interference effects. With added input filtering, an active antenna with a 1-meter whip can provide less IMD because the input filter reduces the likely interfering signals before they have a chance to operate on the preamp input circuitry.

Although the wideband active antenna should not be used with anything longer than a 1-meter whip in areas of high adjacent-channel interference, longer antennas—perhaps up to 10 meters—can be tried in a “quiet” location for operating in the VLF-LF range. However, when using long antennas in the HF region there is an additional interference problem because the antenna is resonant at more than one frequency. One rule to follow here is to keep the length of the antenna less than $\frac{1}{10}$ wavelength at the highest frequency used for a wideband system. Although that is short at the highest frequency, an

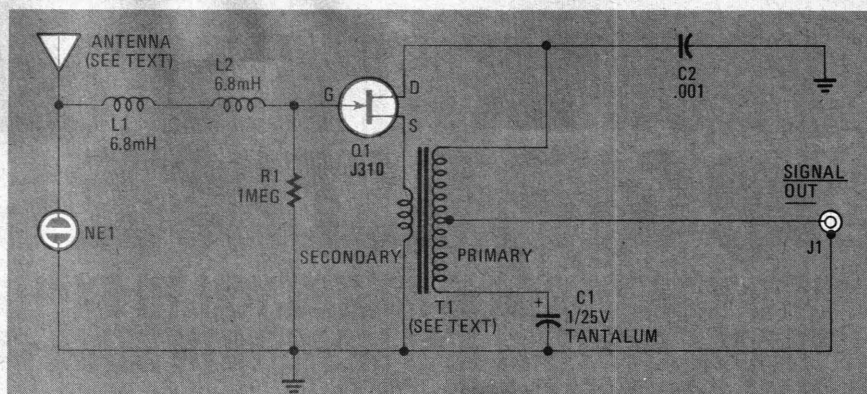


FIG. 4—THE INPUT INDUCTORS and circuit capacitance form a lowpass filter that makes this an amplifier for restricted use in the VLF-LF range.

antenna of that length used with the wideband preamp will perform almost as well as a 48-inch top-loaded vertical connected to a 50-ohm system (as in mobile CB radios at the 27-MHz region). A primary reason for using an active-antenna system is to provide good performance over a wide range with small physical size. Thus, if the antenna is to be used only for the CB range, it would be simpler to use an ordinary CB antenna and avoid all of the wideband problems.

Amplifier circuit—VLF and LF

At frequencies below about 500 kHz, the amplifier circuit is modified to provide input filtering and higher voltage-gain. Figure 4 shows the modified circuit. Two input inductors and the circuit capacitances form a lowpass filter with a cutoff frequency near 450 kHz (see Fig.

5-a). The choice of those inductors is somewhat critical because the preamp's operation depends partly on the resonant frequency of the coils, the distributed capacitance, and the capacitance of the windings to the shield housing. To reduce mutual coupling, the coils are connected in series with their windings opposing each other. Therefore, they still can be mounted close together on a small circuit board with no interstage shield. That arrangement provides at least another 30 dB of attenuation for broadcast-band signals directly at the input to the preamplifier where the problem of intermodulation starts. A single inductor can be used, but it will not provide quite as sharp a cutoff for interference from the AM broadcast band.

The output transformer is an ultra-miniature audio-output transformer with a 200-ohm center-tapped primary and an 8-ohm center-tapped secondary. (We will talk more about that transformer when the series continues.) The output transformer has good response to at least 400 kHz, even though it was originally intended for audio-frequency use. The smaller amount of feedback applied from drain-to-source results in higher voltage gain of about +6 dB at the expense of slightly less power gain, or a higher output impedance when compared to the 1:1 wideband toroid. However, we use the iron core transformer because of its low cost as well as the lowpass output filtering provided.

When used with a 1-meter whip, the VLF-LF version of the active antenna—with an input lowpass filter with about a 450 KHz rolloff—provides higher intercept points with respect to broadcast-band interference (although it is about the same for interference from other frequencies). If you are located in a region free from high-power broadcast-band transmitters, then you can use the preamplifier of Fig. 4 with longer antennas. However, a point is reached with any active system where merely increasing the antenna size does not improve the overall signal-to-noise ratio because the atmospheric noise level increases at the same rate as the signal.

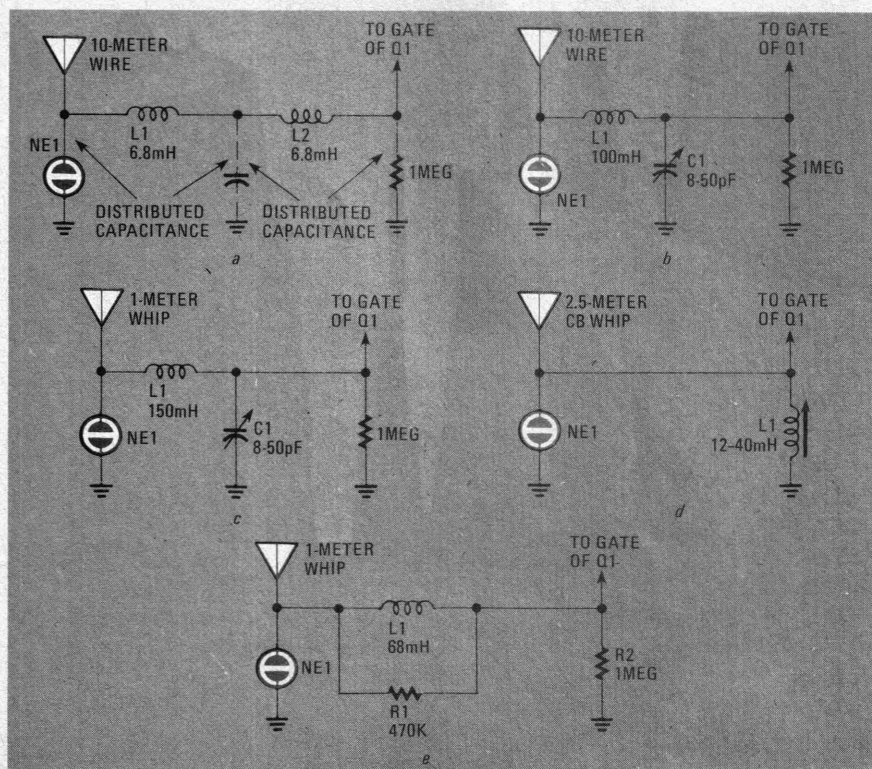


FIG. 5—VARIOUS INPUT NETWORKS for VLF-LF operation can improve performance at particular frequencies or increase the antenna's selectivity.

Resonant input circuit

Figure 5 illustrates various high-impedance input networks for restricted use in the VLF-LF region (such as Loran-C only, or WWVB, or for the 160 kHz–190 kHz experimenters' license-free band). A series inductor with a small input tuning capacitor can be used to further reduce interference and increase the antenna performance. A miniature trimmer-capacitor with a tuning range of 8 to 50 pF placed from the gate to ground, directly across the 1 megohm input resistor (see Figs. 5-b and 5-c) provides a means of tuning the series inductor for a peak at the desired frequency range. The result is a sharp, high-frequency cutoff with a more gradual low-frequency rolloff. The inductor was chosen to be self-resonant (remember, real inductors also have capacitance) at a somewhat higher frequency than the top of the desired tuning range. That technique will work for some pot-core or slug-wound inductors but will usually not work well with large toroids, as they have too much distributed capacitance at VLF. It is also possible to shunt a slug-tuned inductor from the gate to ground (as in Fig. 5-c) but the preamp will then require a larger housing. For that parallel-tuned case, the 1-megohm resistor can be removed because the inductor provides the ground return for the gate. The antenna is then connected directly to the gate terminal with the inductor chosen to resonate with the antenna, input-circuit, and antenna-mount capacitances. The minimum of external tuning capacitance provides the highest Q (most selective) antenna in this application. For DX hunting in the low-frequency experimenters' band (at 180 kHz), a narrowband antenna with a Q of more than 50 can be achieved with a parallel-tuned circuit.

One problem with using a tuned circuit is that it restricts the remote applications of the active antenna. That is because the antenna must be located conveniently so that it can be retuned. However, for covering some fixed frequency (such as

the experimenters' band) the antenna system can be aligned on the bench and then mounted for unattended operation. When tuning those systems, it is advisable to temporarily mount the preamplifier assembly in a fairly clear area (preferably where it will be permanently located) to avoid nearby capacitive coupling, which might detune a very selective system.

One technique for broadbanding a tuned circuit is to place a resistor in parallel with the inductor (See Fig. 5-d). Resistor values in the range of 50K to 500K ohms can help broaden Loran-C systems where a wide bandwidth is necessary.

Traps

Series-connected transmission-line traps tuned to local broadcast-band stations and placed just ahead of the receiver coupler can improve the IMD somewhat and reduce overload or gain-compression problems (see Fig. 6). The tuning capacitors must be isolated from ground and the inductor must be chosen so as to have a reactance greater than 50 ohms at the desired notch frequency. Dual traps are possible. For example, Fig. 6 shows a trap for 970 kHz and another for 1340 kHz connected in series. The combination of input lowpass filters at the antenna and traps at the preamp output can usually provide sufficient attenuation for cases of severe interference in the VLF-LF band from stations in the broadcast band.

A summary of some measurements made with different antennas at 60 kHz for WWVB reception is shown in Table 1. It should be noted that a 2-meter vertical whip is about equivalent in sensitivity to the much larger flat-top antenna. However, the flat top is much more susceptible to noise and interference, even when it is operated with a lowpass filter at the preamp input. The effective-height estimate may not be the same over the entire frequency range. For example, the flat top appears to have an effective height of about 2 meters at 200 kHz but less than 0.9 meters at 60 kHz. That is because of

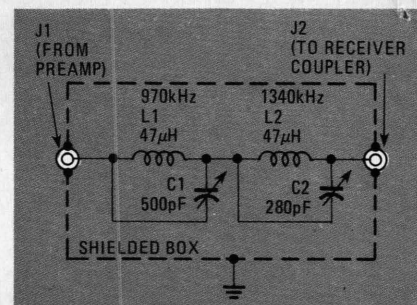


FIG. 6—TRAPS CAN BE USED TO reduce interference from broadcast band stations—in this case from stations at 970 and 1340 kHz.

K—the shielding effect and conductivity of the local ground terrain, which includes all the trees, power lines, and building structures. However, we are still able to operate the antenna even down to the 10.2 kHz Omega frequency with reasonable success and it is used routinely to check GBR on 16 kHz for VLF propagation conditions. (GBR is a high-power military VLF station from Great Britain.) In practice, it is always wise to check for IM effects at the specific frequency range that you plan to use the antenna. Sometimes they are severe but only at relatively narrow frequency ranges usually not in the VLF range.

For general wideband surveillance, the 1-meter whip with an effective height of about 30 cm is the best antenna of all, because it has fewer IM interference effects and less local noise from the power lines.

A general conclusion from all of the experiments is that the local environment and the ground-conductivity effects of nearby structures are the most important factors in determining antenna sensitivity. Small changes in antenna location can produce remarkable differences in the antenna's performance.

Another observation is that the best location for a short whip is invariably up high in the clear. (That can especially be seen in aircraft applications where a very short vertical whip is used with remarkably good performance.)

Low-frequency experimental radio station operators have reported good results in mobile operation with reception of 160 to 190 kHz signals using 2.5-meter CB whips and parallel-tuned input networks. We have conducted similar experiments with Omega and Loran-C receivers in mobile vehicles where the only problems were those of shielding from buildings or when driving under bridges or near power lines. An additional problem in mobile operations is harmonic radiation from the vehicle's AC alternators.

When we continue this series, we will discuss construction details and include printed-circuit board layouts for the active antenna preamplifier and receiver coupler. We will also discuss how to bench test the preamp, and how to mount the system.

R-E

TABLE 1

Parameter		Whip 1	Whip 2	Flat Top
Physical Height	(h_m)	1m	2m	10m
Antenna Capacitance	(C_a)	10pF	20pF	118pF
Fixed Capacitance	($C_m + C_a$)	15pF	15pF	15pF
Voltage Gain at Preamp	(A_v)	1 (0dB)	1 (0dB)	2 (+6dB)
Estimated Ground Coupling Effect	(K)	0.7	0.7	0.05
Effective Height	(h_e)	0.28m	0.80m	0.88m
WWVB Reading on YAESU FRG-7700		S6	S9	S9+
Estimated E-field for WWVB (from NBS chart)	(E_i)	150 μ V-per-m	150 μ V-per-m	150 μ V-per-m
Output S+N for ($E_o = E_i \times h_e$)		42 μ V	120 μ V	132 μ V
60kHz WWVB at Preamp				
Estimated (S+N)/N during 60 Hz "quiet hours"		+10dB	+20dB	+20dB
Overall Noise Rating		good	fair	poor
IM Distortion Rating		fair	poor	very poor

EXPERIMENTING WITH MAGNETIC SENSORS

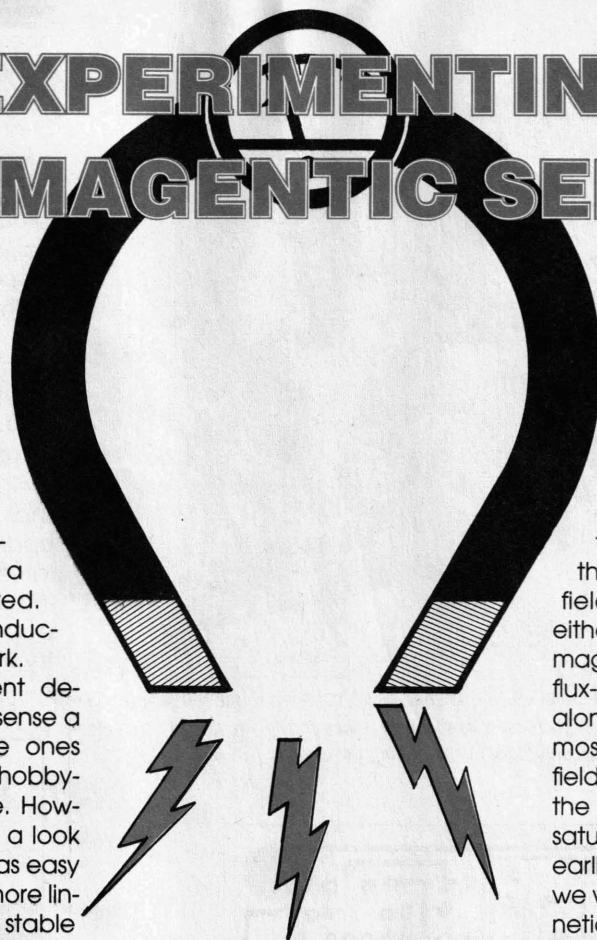
Magnetic fields are all around us. The Earth itself produces a magnetic field, which is why compasses work. Anytime an electrical current flows in a conductor, a magnetic field is generated. That is why transformers, inductors, and radio antennas work.

There are several different devices that could be used to sense a magnetic field. One of the ones most familiar to electronics hobbyists is the Hall-effect device. However, in this article we'll take a look at a magnetic sensor that is as easy to use, but is more sensitive, more linear, and more temperature stable than typical Hall-effect devices. And just like Hall-effect devices, it can be used to make a variety of instruments, including magnetometers and gradiometers.

For those unfamiliar with them, magnetometers are used in a variety of applications in science and engineering. One high-tech magnetometer is used by Navy aircraft to locate submarines. Radio scientists use magnetometers to monitor solar activity. Archeologists use magnetometers to locate buried artifacts, while marine archeologists and treasure hunters use the devices to locate sunken wrecks and sunken treasure.

Anyway, the device we will be exploring is called a "flux-gate magnetic sensor." The device, in essence, is basically an over-driven magnetic-core transformer in which the "transducible" event is the saturation of the magnetic material. These devices can be made very small and compact, yet will still provide reasonable accuracy.

The most simple form of flux-gate magnetic sensor is shown in Fig. 1A.



Flux-gate magnetic sensors can provide superior performance to other popular alternatives, and a new device makes them easier than ever to use.

JOSEPH J. CARR

It consists of a nickel-iron rod used as a core, wound with two coils. One coil is used as the excitation coil, while the other is used as the output or sensing coil. The excitation coil is driven with a squarewave (see Fig. 1B) with an amplitude high enough to saturate the core. The current in the output coil will increase in a linear manner so long as the core is not saturated. But when the saturation point is reached, the inductance of the coil drops and the current rises to a level limited only by

the coil's other circuit resistances.

If the sensor of Fig. 1A were in a magnetically pure environment, then the magnetic field produced by the excitation coil would be the end of the story. But there are magnetic fields all around us, and these either add to or subtract from the magnetic field in the core of the flux-gate sensor. Magnetic field lines along the axis of the core have the most effect on the total magnetic field inside the core. As a result of the external magnetic fields, the saturation condition occurs either earlier or later than would occur if we were only dealing with the magnetic field of interest. Whether the saturation occurs early or later depends on whether the external field opposes or reinforces the intended field.

A better solution is shown in Fig. 2. In this version of the flux-gate sensor, there are two independent cores, each of which has its own excitation winding. A common pick-up winding serves both cores. The excitation coils are wound in a series-opposing manner so that the induction generated in the cores precisely cancels each other if the external field is zero. However, in the presence of an external field, pulses are produced in the pick-up coil that can be integrated in a low-pass filter to produce a slowly varying DC signal that is proportional to the applied external magnetic field.

Toroidal-Core Flux-Gate Sensor.

The flux-gate sensors that use a linear or straight core suffer from two main problems. First, the desired signal is small compared with the signal on which it rides, so is difficult to discriminate properly. Second, there must be a very good match

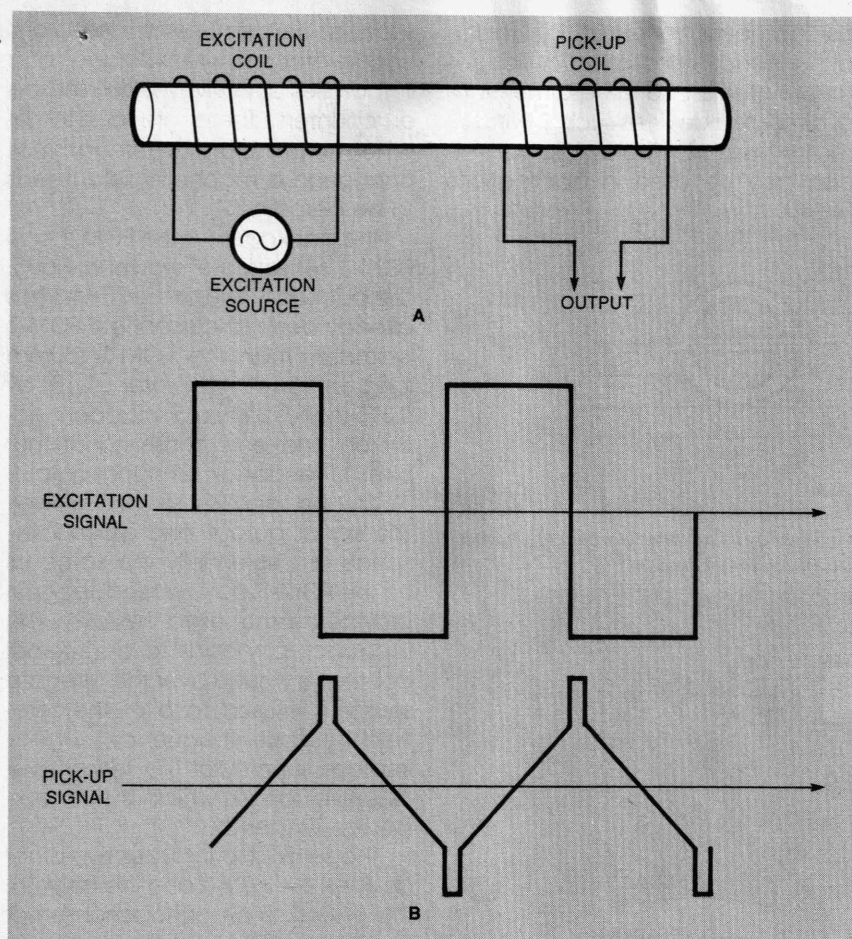


Fig.1. Here's the flux-gate magnetic sensor in its simplest form (A) and a sample excitation signal and the output that it produces (B).

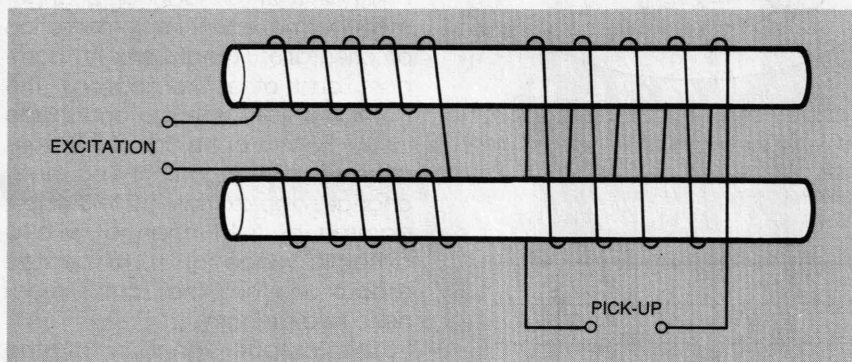


Fig. 2. While the simple circuit of Fig. 1 would work under ideal conditions, because we are constantly surrounded by all kinds of magnetic fields, the dual-core design shows here would produce superior results.

between the cores and the excitation winding segments on each winding. While those problems can be overcome, it becomes expensive, limiting the design's popularity.

A better solution is to use a toroidal or "doughnut"-shaped magnetic core. That type of core relieves the problem of picking off small signals in the presence of large offset components. It also reduces the drive levels required from the

excitation source.

In the toroidal-core flux-gate sensor, we can get away with using a single excitation coil wound over the entire circumference of the toroidal core (see Fig. 3). The pick-up coil is wound over the outside diameter of the core.

Another advantage of the toroidal-core version of the flux-gate sensor is that a pair of orthogonal (i.e. right angle) pick-up cores can

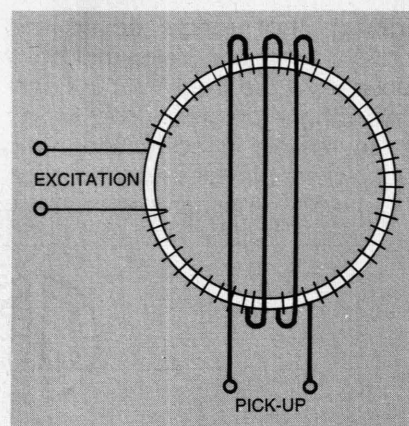


Fig. 3. Using a toroid core overcomes some of the problems and limitations inherent in the straight-core design

be installed that will allow null measurements to be made. Figure 4 shows the orientation of the toroid-core flux-gate sensor as a function of sensitivity. The maximum sensitivity occurs when the magnetic H-field is orthogonal to the pick-up coil, while minimum sensitivity occurs when the pick-up coil and H-field are aligned with each other. As you can see, when there are two pick-up coils at right angles to each other, one will be at maximum sensitivity when the other is at a null (minimum sensitivity).

A Practical Flux-Gate Sensor. A compact and reasonably low-cost line of flux-gate sensors, designated FGM-x, is made by Speake & Co. Ltd (Elvicta Estate, Crickhowell, Powys, Wales, UK), and distributed in the United States by Fat Quarters Software (24774 Shoshonee Drive, Murrieta, CA 92562; Tel: 909-698-7950, Fax: 909-698-7913; e-mail: 73162,2364@compuserve.com). The FGM-3 device is the one that I experimented with when preparing this article. It is a 62mm-long by 16mm-diameter (2.44 by 0.63 inch) device. Like all the devices in the line, it converts the magnetic field strength to a signal with a proportional frequency. One of the things I found fascinating about the FGM-3 is that a set of only three leads provides operation: Red: +5 VDC (power), Black: 0 Volts (ground), and White: output signal (a squarewave whose frequency varies with the applied field).

The magnetic detection rating of the device is ± 0.5 oersted (± 50

μ tesla). That range covers the earth's magnetic field, making it possible to use the sensor in Earth-field magnetometers. Using two or three sensors in conjunction with each other provides functions such

as compass orientation, three-dimensional orientation measurement systems, and three-dimensional gimballed devices such as virtual-reality helmet display devices. It can also be used in applications

such as ferrous metal detectors, underwater shipwreck finders, and in factories as conveyor-belt sensors or counters. There are a host of other applications where a small change in a magnetic field needs to be detected.

The packages for the FGM-1 and FGM-3 sensors are shown in Fig. 5. The package style of the FGM-3 has already been discussed. The FGM-1 is smaller than the FGM-3 (30mm long by 8mm diameter; 1.18 by 0.315 inches). It has a small connector on one end consisting of four pins: 1) feedback; 2) signal output; 3) ground, and 4) +5 VDC power. The signal, output, and ground terminals are essentially the same as on the FGM-3, but the feedback pin provides some extra flexibility. The feedback pin leads to an internal coil that is wound over the flux-gate sensor. It is used to alter the zero-field output frequency, or to improve linearity of the sensor over its entire range, which is ± 0.7 oersted ($\pm 70 \mu$ tesla).

The series also includes two other devices, the FGM-2 and the FGM-3h. The FGM-2 is an orthogonal sensor that has two FGM-1 devices on a circular platform at right angles to one another. That orthogonal arrangement permits easier implementation of orientation measurement, compass, and other applications. The FGM-3h is the same size and shape as the FGM-3, but is about 2.5 times more sensitive. Its output frequency changes approximately 2 to 3 Hz per gamma of field change, with a dynamic range of ± 0.15 oersted (about one-third the Earth's magnetic field strength).

The output signal in all the devices in the FGM series is a +5 volt (TTL-compatible) pulse whose period is directly proportional to the applied magnetic field strength. This relationship makes the frequency of the output signal directly proportional to the magnetic field strength (Fig. 6). The period varies typically from 8.5μ s to 25μ s, or a frequency of about 120 kHz to 50 kHz. For the FGM-3 the linearity is about 5.5 percent over its ± 0.5 oersted range.

The response pattern of the FGM-x series sensors is shown in Fig. 7. It is a "figure-8" pattern that has major lobes (maxima) along the axis of the

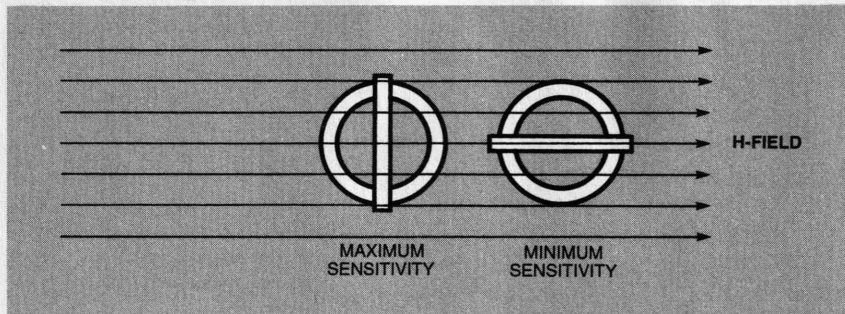


Fig. 4. The sensitivity of the toroid-wound sensor is affected by its orientation within the magnetic field. That property can be put to good use in a variety of sensor applications.

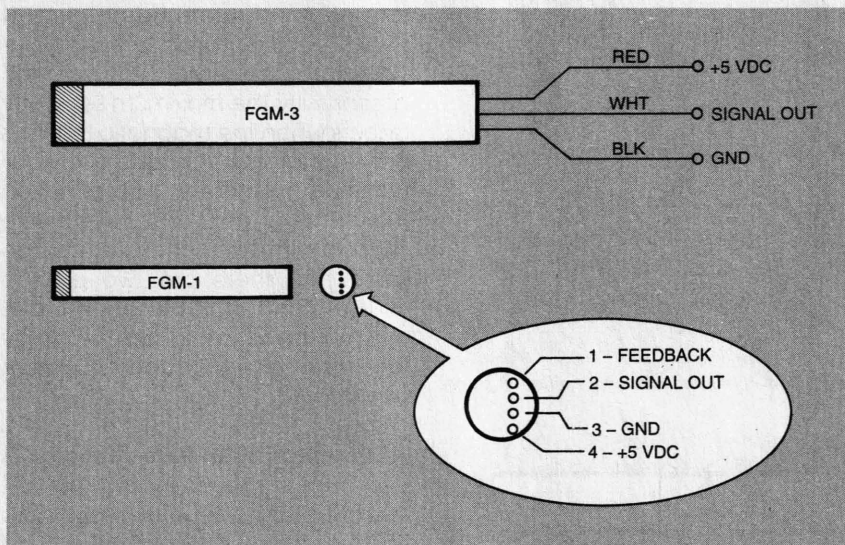
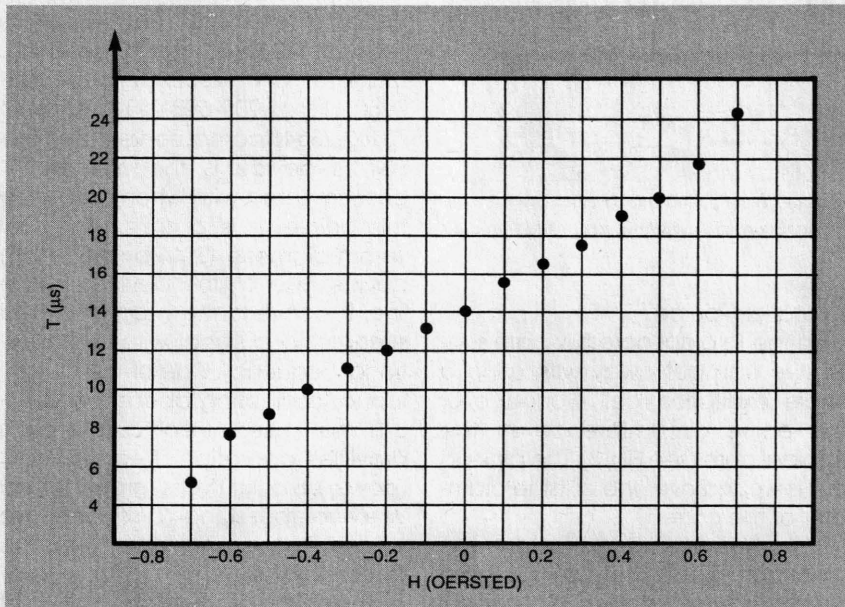


Fig. 5. Here are the pinouts of the FGM-1 and FGM-3, two members of a family of practical flux-gate sensors made in Wales, UK by Speake & Co. Ltd.



58 Fig. 6. The output of the FGM devices is directly proportional to the magnetic-field strength.

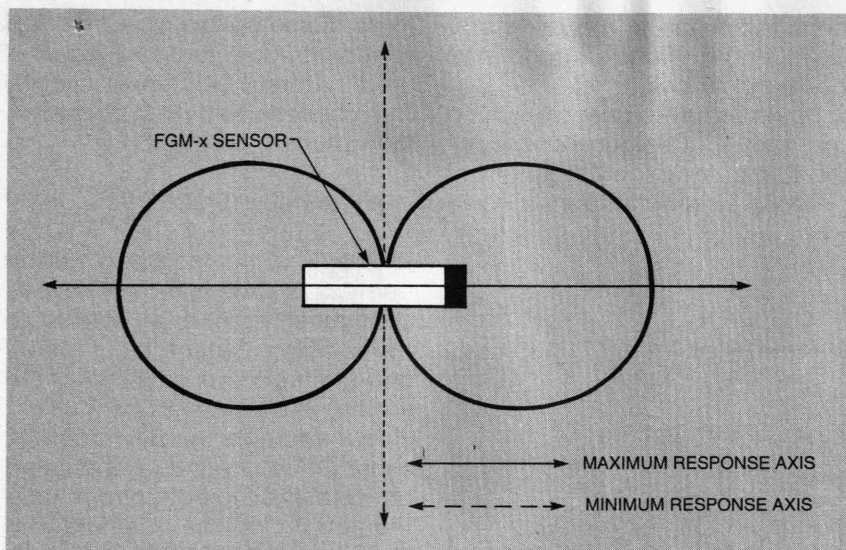


Fig. 7. The figure-eight sensitivity pattern for the FGM-x devices makes them most sensitive along their long axis, and least sensitive at right angles to it.

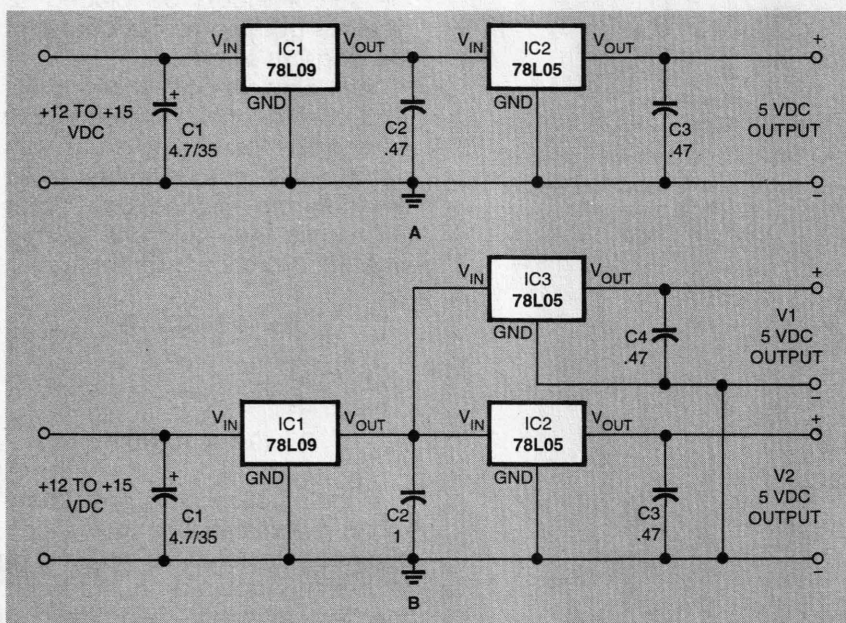


Fig. 8. As ripple will adversely affect the operation of the device, a double-regulated DC power supply (A) should be used to power the device. If other circuits are also to be driven from the supply, use the version shown in B.

sensor, and nulls (minima) at right angles to the sensor axis. This pattern suggests that for any given situation there is a preferred direction for sensor alignment. The long axis of the sensor should be pointed towards the target source. When calibrating or aligning sensor circuits, it is common practice to align the sensor along the east-west direction in order to minimize the effects of the Earth's magnetic field.

Powering the Sensors. The FGM-x series of flux-gate magnetic sensors operates from +5 volts DC, so it is

compatible with a wide variety of analog and digital support circuitry. As is usual for any sensor, you will want to use only a regulated DC power supply for the FGM-x devices. In fact, the manufacturer recommends that double-regulation (Fig. 8A) be used. Ripple in the DC power supply can cause output frequency anomalies, and those should be avoided. In the circuit of Fig. 8A, an unregulated +12 to +15 VDC input potential is applied first to a 9 volt 78L09 or 78M09 three-terminal IC voltage regulator (IC1). That produces a +9 volt regulated potential

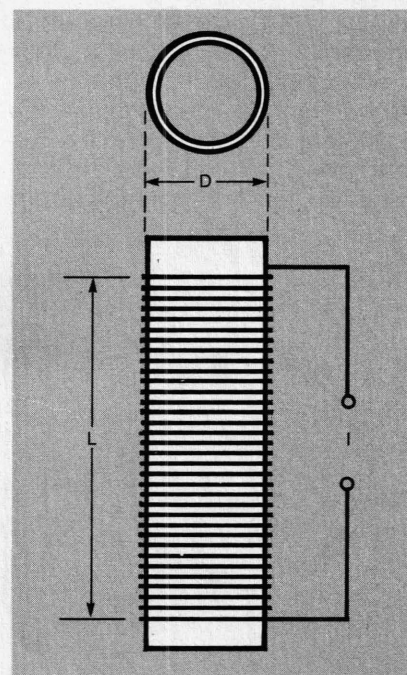


Fig. 9. To calibrate the FGM devices, a solenoid-wound coil like this one could be used.

that is then applied to the input of a 78L05 or 78M05 device (IC2). The second regulator reduces the +9 volts from IC1 to the +5 volts needed for the FGM-x sensors.

When other digital devices are being powered from the same DC power supply, it is prudent to provide a separate DC source for the FGM-x sensors. In Fig. 8B we see a circuit that would accomplish that task. There are two separate +5-volt DC outputs, labeled V1 and V2. Both are derived from 78L05 devices that are powered from a single 78L09. Care must be taken to not exceed the maximum current limits of IC1, especially if the same size IC voltage regulators are used for all three (IC1-IC3). One of the +5 VDC sources, either V1 or V2, can be used for powering the FGM-x device, while the other powers the rest of the circuitry.

Calibrating the Sensors. The FGM-x devices are not precision instruments straight out of the box, but can be calibrated to a very good level of accuracy. The calibration chore requires you to generate a precise magnetic field in which the sensor can be placed. One way to generate well-controlled and easily measured magnetic fields is to build a coil and pass a DC current

through it. If the sensor is placed at the center of the coil (inside), then the magnetic field can be determined from the coil geometry, the number or turns of wire and the current through the coil. There are basically two forms of calibrating coil

found in the various magnetic sensor manuals: solenoid-wound and the Helmholtz pair.

Figure 9 shows a solenoid-wound coil: a coil that is wound on a cylindrical form in which the length of the coil (L) is greater than or equal

to its diameter. This type of coil is familiar to radio fans because it is used in many L-C tuning circuits. The magnetic field (H) in oersteds is found from:

$$H = (4\pi NI)/(10 \text{ SQRT } (L^2 + D^2))$$

Where: H is the magnetic field in oersteds, N is the number turns-per-centimeter (t/cm) in the winding, I is the winding current in amperes, and D is the mean diameter of the winding in centimeters (cm)

The winding is usually made with either 24- or 26-gauge enameled or *Formvar*-covered copper wire. The length of the solenoid coil should be at least twice as long as the sensor being calibrated, and the sensor should be placed as close as possible to the center of the long axis of the coil.

The Helmholtz coil is shown in Fig. 10. It consists of two identical coils (L_1 and L_2) mounted on a form with a radius of R , and a diameter of $2R$. The coils are spaced one radius ($1R$) apart. The equations for that type of calibration assembly are:

$$H = (0.8991NI)/R$$

and,

$$B = (9.1 \times 10^3 NI)/R$$

In the practical case, one usually knows the dimensions of the coil, and needs to calculate the amount of current required to create a specified magnetic field. We can get that for the Helmholtz pair by rearranging the first equation of the pair to:

$$I = (RH)/(0.8991N)$$

The coils are a little difficult to wind, especially those of large diameter (e.g. 4 inches). One source recommends using double-sided tape (the double-sticky stuff) wrapped around the form where the coils are to be located. As the wires are laid down on the form they will stick to the tape, and not dither around.

The above equations, plus a lot of magnetic theory and calibration suggestions are found in *The Mag-*

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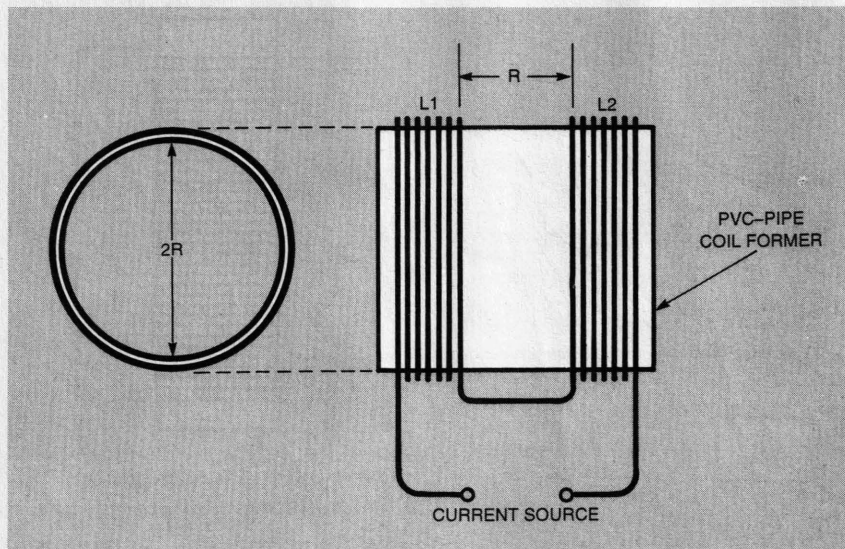
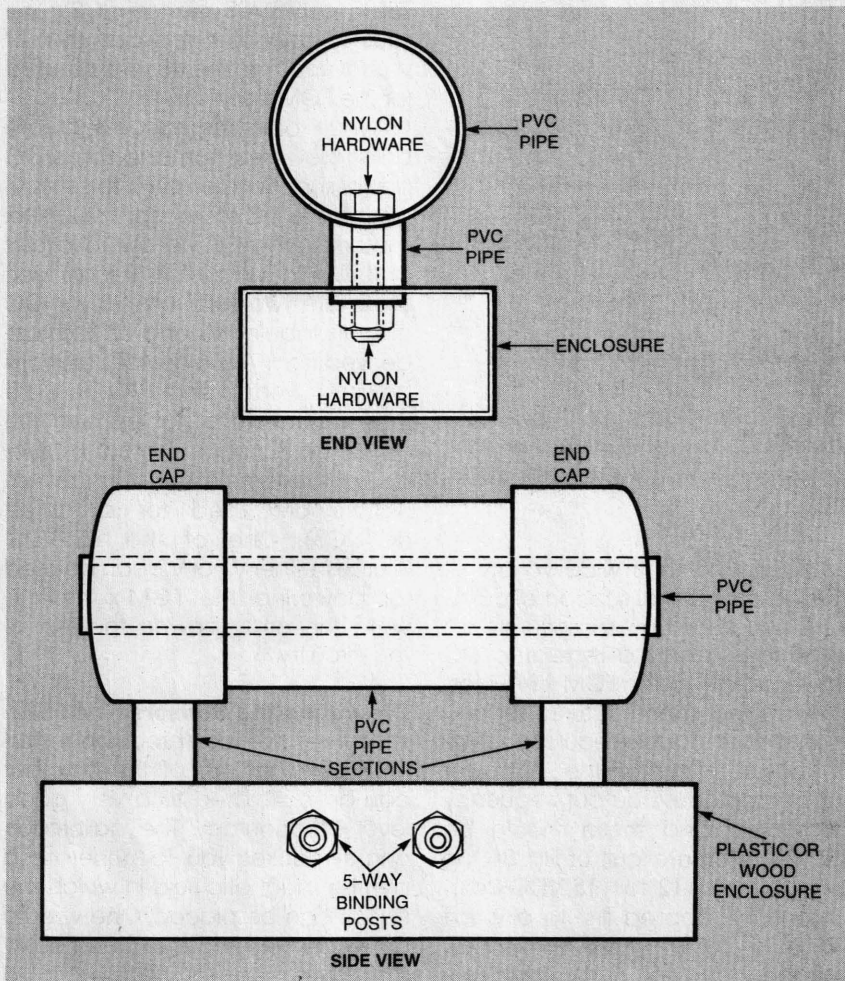


Fig. 10. A Helmholtz-pair calibration coil could also be used, though these can be difficult to wind.



60 Fig. 11. Here's a simple coil-mounting assembly that can be used in your calibration setup.

Plan Your Escape



Fire Can Happen Any Time!

And if it does, every second counts. Will you and your family know what to do?

- Practice an escape plan from each room in the house. Feel your way out with your eyes closed.
- Teach your family to stop, drop to the ground and roll if their clothes catch on fire.
- Keep the fire department's number by the phone. By your bed, have a flashlight to help you see and whistle to alert your family.
- Install smoke detectors on every level of your house. Test them monthly, and change the batteries at least once a year.



Remember, fire can happen anytime, so make plans for a surprise visit.



A message from the U.S. Fire Administration

ABOUT THE AUTHOR

Albert Lozano-Nieto, Ph.D. is an Assistant Professor of Engineering at the Penn State University, Wilkes-Barre Campus, where he is responsible for the Associates Degree in Biomedical Engineering Technology. Dr. Lozano-Nieto has worked in academia as well as with private and government industry, especially in the subjects of Bioengineering and Electromagnetic Compliance of electronic instruments. He has been involved in research projects with the European Space Agency, designing equipment to monitor physiological parameters for in-flight astronauts. Dr. Lozano-Nieto can be reached by email at: AXL17@psu.edu, by phone at (717) 675-9245, or by mail at P.O. Box PSU, Lehman, PA 18627.

have an interest in learning how things work. You get extra credit if, as a child, you loved to take things apart. And you get even more extra credit if you were able to put things back together without missing pieces or having extra ones!

Most of the academic institutions that train biomedical engineering technicians offer a two-year college program that leads to an Associate Degree in Technology. The main emphasis of these college programs is on the electronic aspects of medical instruments: the safety issues, how physiological measurements are made on the human body, and how the medical equipment works. Also covered are related issues and technologies such as computers, telecommunications, mechanics, and fluids; education in these disciplines is important as they all converge in medical instrumentation.

If you are interested in pursuing a degree in this field, contact the colleges near where you live to obtain more detailed information. There are not a lot of universities and colleges that offer this degree, so you might need to do a little bit of research on your own. In any case, you can contact me (see the "About the Author" sidebar), and I will help you find a suitable institution near you. Ω

MAGNETIC SENSORS

(continued from page 60)

netic Measurements Handbook by J.M. Janicke (\$20, Magnetic Research Press, 122 Bellevue Avenue, Butler, NJ, 07405).

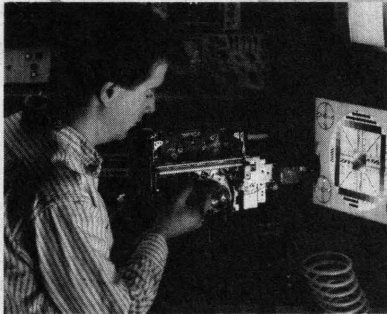
Figure 11 shows a type of assembly that can be used for either the solenoid or Helmholtz coil. I first saw this type of assembly in a college freshman physics laboratory about 25 years ago. It consists of a PVC pipe section used as the coil former. End caps on the coil former also serve as mountings. The mounts at either end consist of smaller segments of PVC pipe and nylon (non-magnetic) hardware fasteners. Another segment of PVC pipe, of much smaller diameter than the coil former, is passed through the former from one end-cap to the other, such that its ends protrude to the outside. This pipe forms a channel into which the sensor can be placed. The base is a plastic or wooden box (again, non-magnetic materials). One thing nice about that type of assembly is that the sensor is always in approximately the same position in the coil, and close to the center of the field.

Conclusion. Now that we've established a base of understanding of flux-gate sensors, it is a simple matter to put that knowledge to use to build some practical magnetometer and gradiometer projects. That will be the focus of another article, which will appear shortly. In the meantime, if you have any question or comments on what we've discussed here, you can contact me directly at PO Box 1099, Falls Church, VA 22041, or via Internet E-mail at carrjj@aol.com. Also, before we wrap up, I'd like to acknowledge the assistance of Richard Noble of Speake & Co. Ltd, and Erich Kern of Fat Quarters Software, the USA distributor of Speake sensors. Ω



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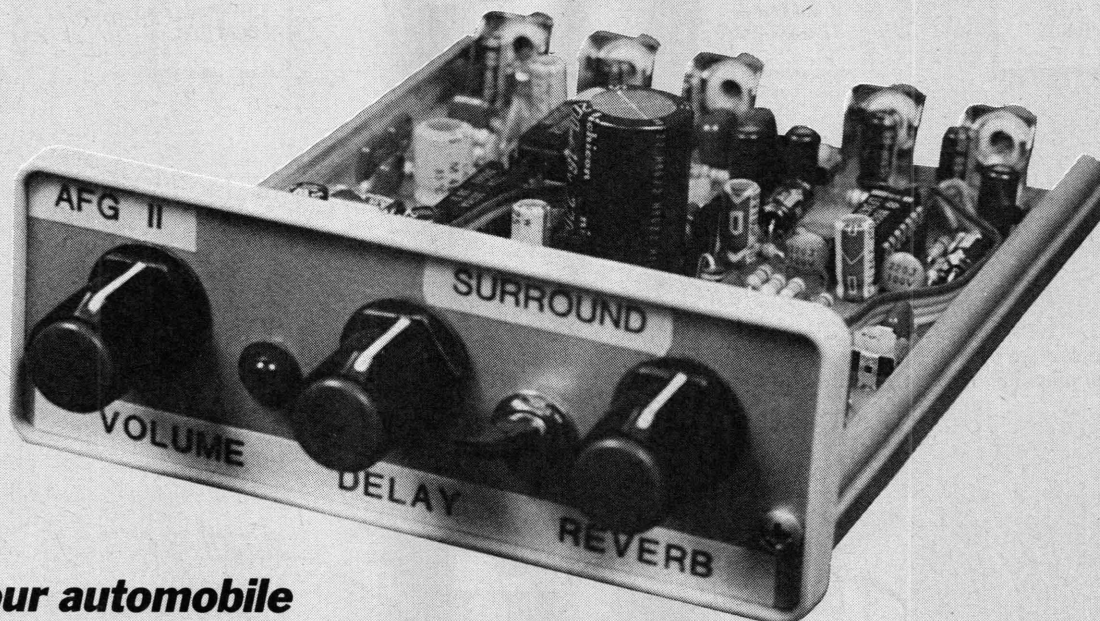
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WAE68

BUILD THE AFG II



Turn your automobile into a concert hall on wheels with the Acoustic Field Generator II.

TOD T. TEMPLIN

WAY BACK IN JANUARY 1990, WE PUBLISHED plans describing the construction of the Acoustic Field Generator. That circuit, the AFG for short, will recover natural ambience and reflections hidden within any stereo audio signal. In other words, it will reconstruct a three-dimensional acoustic field by reproducing both direct and delayed indirect sound waves through a set of four speakers surrounding the listener. The AFG can also decode specially encoded surround sound material, and thus was intended primarily for use in a home theater system.

Many builders of the AFG were so impressed with it that they requested information on how to modify the circuit to operate from a 12-volt power supply for use in their automobiles. It seems that a good number of audiophiles are looking for a circuit to enhance the acoustic properties of their automotive sound systems.

The circuitry of the original AFG included a bipolar power supply and was not readily adaptable to 12-volt operation. For that reason, an entirely new circuit, the AFG II, has been designed for use in a car. However, the AFG II can also be powered from a 12-volt DC wall transformer, and can therefore become part of any audio system.

Description

The AFG II is a true stereo circuit. It's designed to be placed in the line-level signal path of your car's audio system between the output of the receiver and the input of the equalizer or power amplifier, which drives the rear speakers of your vehicle. Figure 1 is a block diagram of a typical system. The AFG II uses a pair of bucket-brigade devices to delay the sound that is fed to the rear speakers. The delay (in relation to the front speakers) can be adjusted from about 5 to 35 milliseconds. A feedback control adds variable reverberation (echo) to the delayed signals.

In addition to its simple delay mode, the AFG II has a differential mode in which it decodes any surround-sound or ambience signals present. Volume and reverberation level are controlled by an electronic volume control IC. As in the original AFG, an electronic crossover is provided to drive an external power amplifier for a subwoofer system.

The entire unit is built on a single-sided 4- by 5-inch PC board with all controls and input/output jacks mounted on it. The completed PC board slides into an attractive commercial enclosure, so the finished project is quite compact, looks professional, and is easily mounted in most vehicles.

Circuitry

Figure 2 shows a block diagram of the AFG II circuit. Note that the circuit consists of two identical audio channels which share a common power supply and bucket-brigade clock generator. Next refer to the schematic

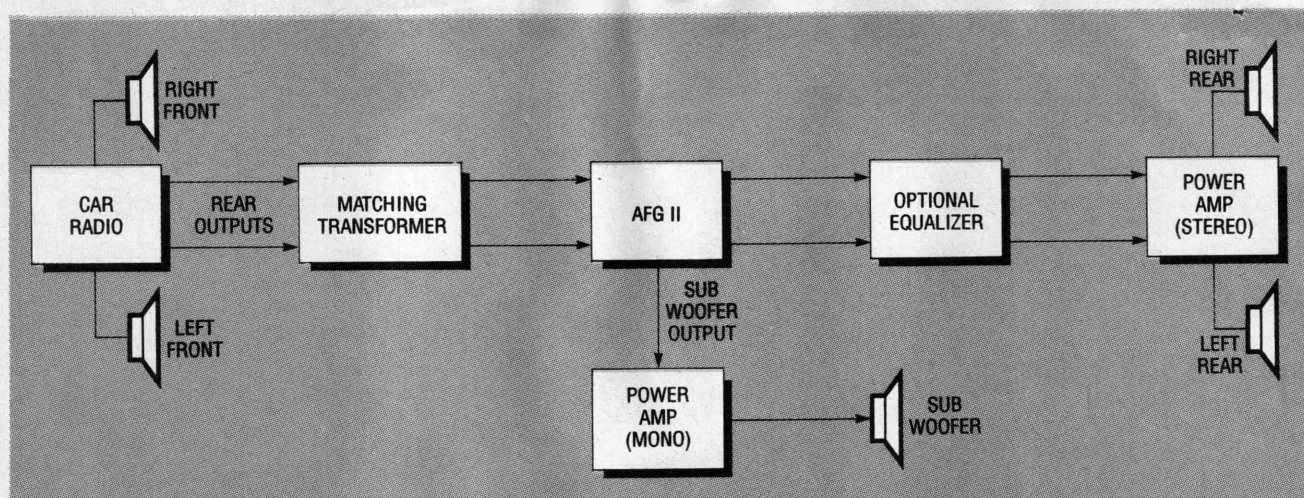


FIG. 1—THE AFG II IS DESIGNED to be placed in the line-level signal path of your car's audio system between the outputs of the receiver and the inputs of the equalizer and power amplifier, which drives the rear speakers.

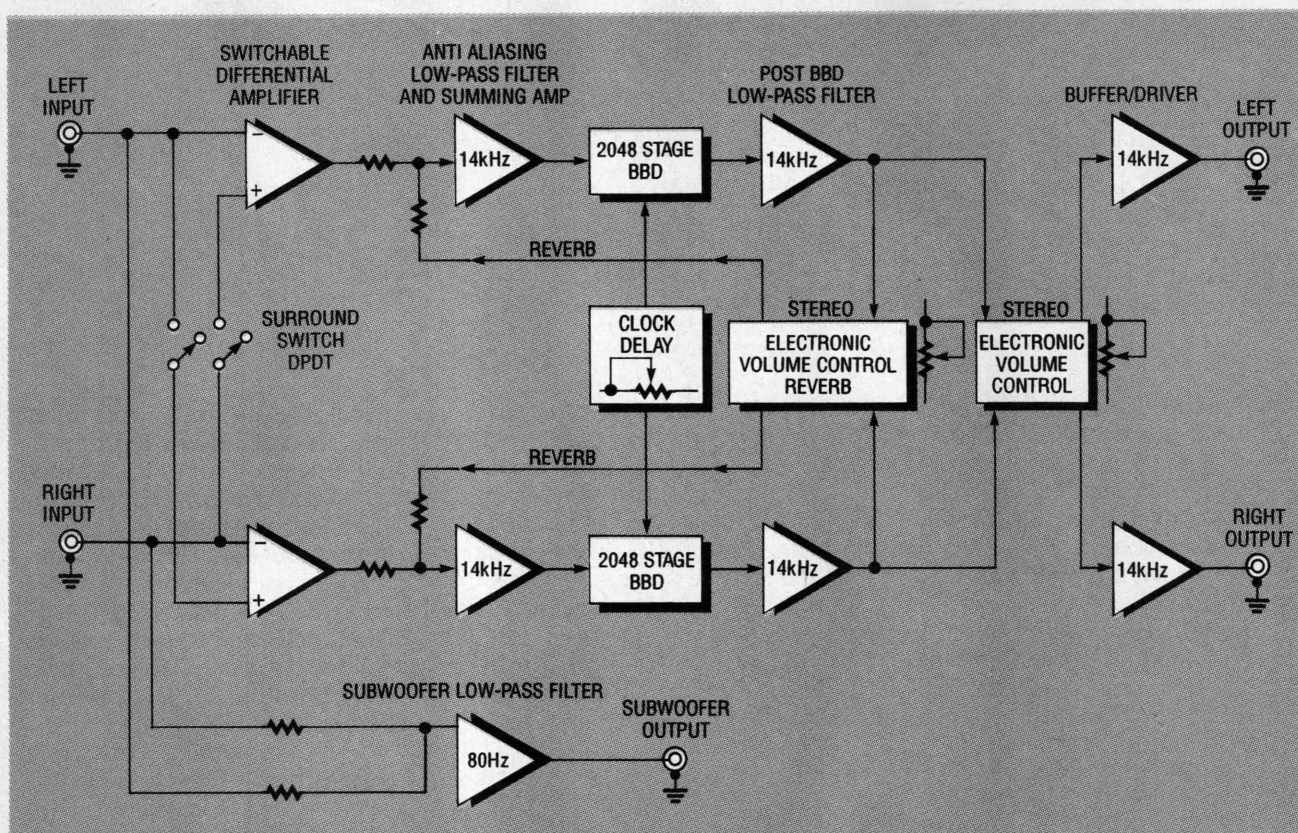


FIG. 2—AFG II BLOCK DIAGRAM. The circuit consists of two identical audio channels that share a common power supply and bucket brigade clock generator.

diagram shown in Fig. 3. An input buffer amplifier is made up of IC1-a and IC1-d. With SURROUND SELECT switch S1 open, IC1-a and -d act as simple inverting amplifiers, and have no effect on the input signal other than providing some gain and a low source impedance for driving the filter section of the following stage.

When S1 is closed, each op-amp becomes a differential am-

plifier because its non-inverting input is cross-connected, via R3-C6 and R4-C5, to the opposite audio channel of its inverting input. When operating in the differential mode, any audio information that is common to both channels (i.e., mono) is canceled from the output. Only the difference component (L-R and R-L) of the original signal remains. Operating the input amplifier stage in the differen-

tial mode decodes any ambience or surround information present in the incoming signal. Trimmers R73 and R74 provide a means for adjusting the overall circuit gain, thus compensating for the varying source signal levels of different components likely to be used with the AFG II.

When op-amps are operated from a single-ended power supply, their non-inverting inputs

should be biased at the same DC voltage as the inverting inputs; usually one half the supply voltage. To satisfy that requirement in our circuit, resistors R21 and R22 form a fixed voltage divider. Because the resistors have equal value, the divider splits the supply voltage in half. Capacitor C17 acts as a filter and ensures that no audio signal or noise is coupled to the non-inverting inputs of the op-amps. All of the op-amps in the entire circuit are biased in this manner. Op-amps IC1-b and IC1-c form a pair of anti-aliasing filters for the delay section (see Fig. 4).

The heart of the AFG II delay section consists of a pair of Panasonic MN3008 2048-stage analog bucket-brigade devices, or BBD (one for each channel), and a shared MN3101 two-phase variable-frequency clock-pulse generator. Each element in the BBD is one of 2048 series-connected stages. Each stage consists of a small capacitor

that stores an electric charge, and a tetrode transistor switches that charge from one capacitor to the next. Consider each of the 2048 capacitors to be a bucket in a fireman's brigade. A signal presented at the input to the BBD is transferred down the line of capacitors in the same manner in which firemen in a bucket brigade transfer a pail of water from one man to the next. The speed at which the signal is transferred down the line of buckets is controlled by the frequency of the clock signal applied. The more slowly the clock runs, the longer it takes for the signal to travel through the circuit and reach the far end.

When using delay lines, one must contend with a nasty property known as "aliasing." In this case, aliasing means "false signal." When the frequency of a signal applied to the input of a BBD is allowed to become higher (i.e., shorter) than one half of the clock pulse fre-

quency, the sample time duration becomes longer than the actual frequency being sampled. Put another way, the amplitude of that high-frequency signal has a value that changes during the sample time; thus the value stored in the capacitor is not an accurate representation of that instant in time, and the output of the BBD is not "true." To avoid aliasing and the resulting distortion, a low-pass filter is placed ahead of the BBD to limit the input-signal frequency to one half or less of the lowest clock frequency used. The low-pass filter used for this purpose is referred to as an anti-aliasing filter.

Another property of BBD's is that clock phase one drives all of the odd number stages, and clock phase two drives the even number stages. When the signal reaches the end of the line, the output of the last odd stage must be combined with the output of the last even stage to

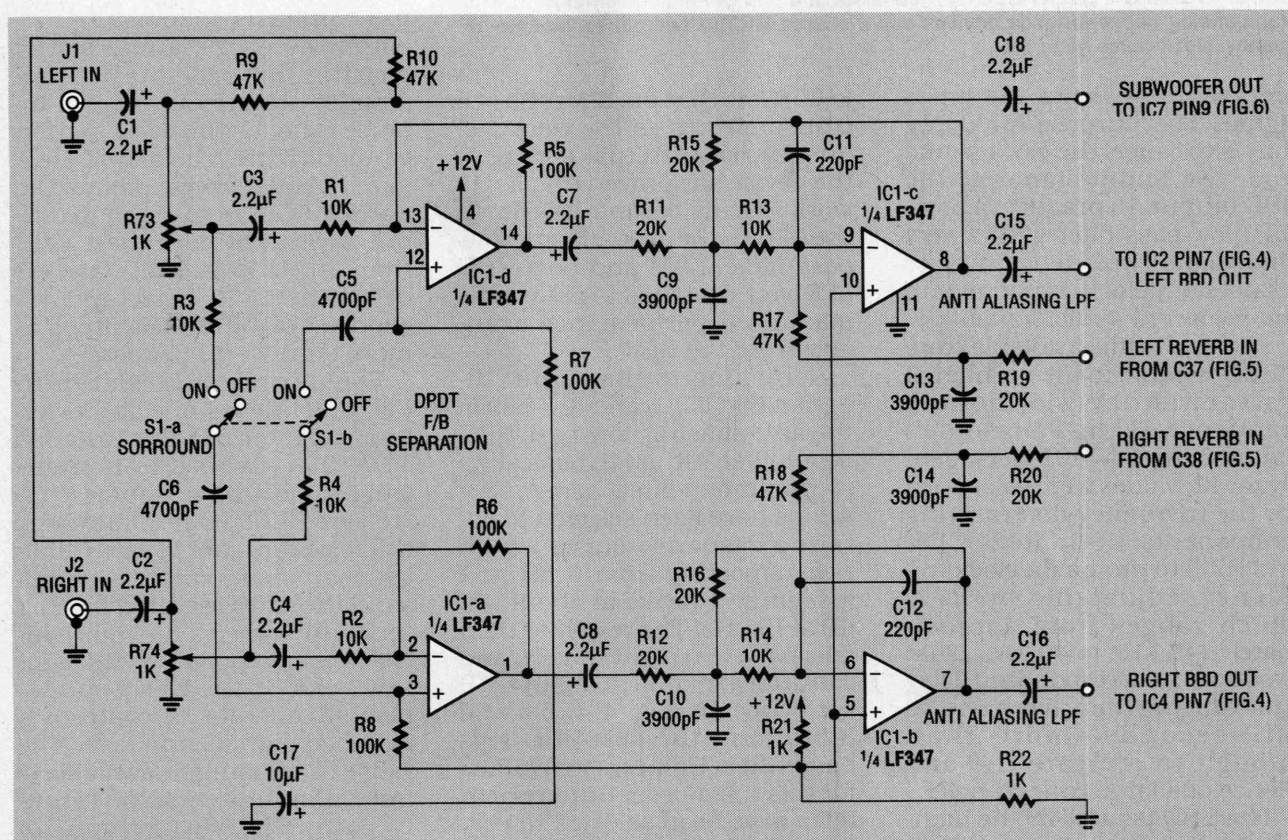


FIG. 3—WITH SWITCH S1 OPEN, IC1-a and IC1-d act as simple inverting amplifiers. With S1 closed, each op-amp becomes a differential amplifier because its non-inverting input is cross-connected to the opposite audio channel of its inverting input. The differential mode "decodes" any ambience or surround information.

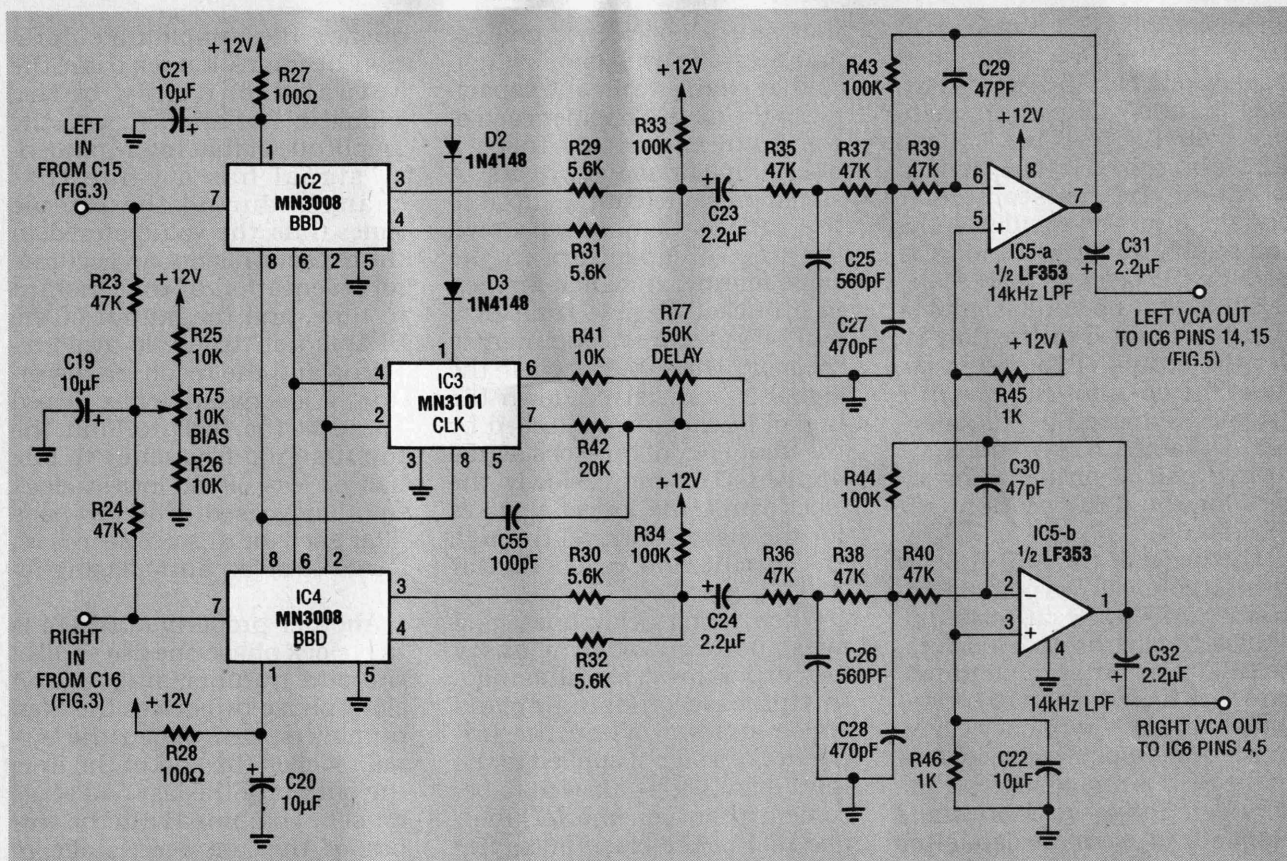


FIG. 4—THE AFG II DELAY SECTION consists of a pair of Panasonic MN3008 2048-stage analog bucket-brigade devices and a shared MN3101 two-phase variable-frequency, clock-pulse generator.

properly reconstruct the input signal. The purpose for doing that is to cancel the clock signal from the audio signal at the BBD output. In practice, a post-BBD low-pass filter with a very sharp cutoff frequency further reduces any clock component in the recovered audio signal.

The delay time available from a BBD is equal to the number of stages divided by twice the clock frequency. Using Panasonic's data for the MN3101 clock generator IC, values were calculated for the frequency-determining components (R41, R42, C55, and R77) to produce a clock frequency, adjustable by R77, which ranges from approximately 145 kHz to 40 kHz. That produces a corresponding delay time ranging from about 5 to 30 milliseconds—more than enough to recreate the ambience of a large concert hall.

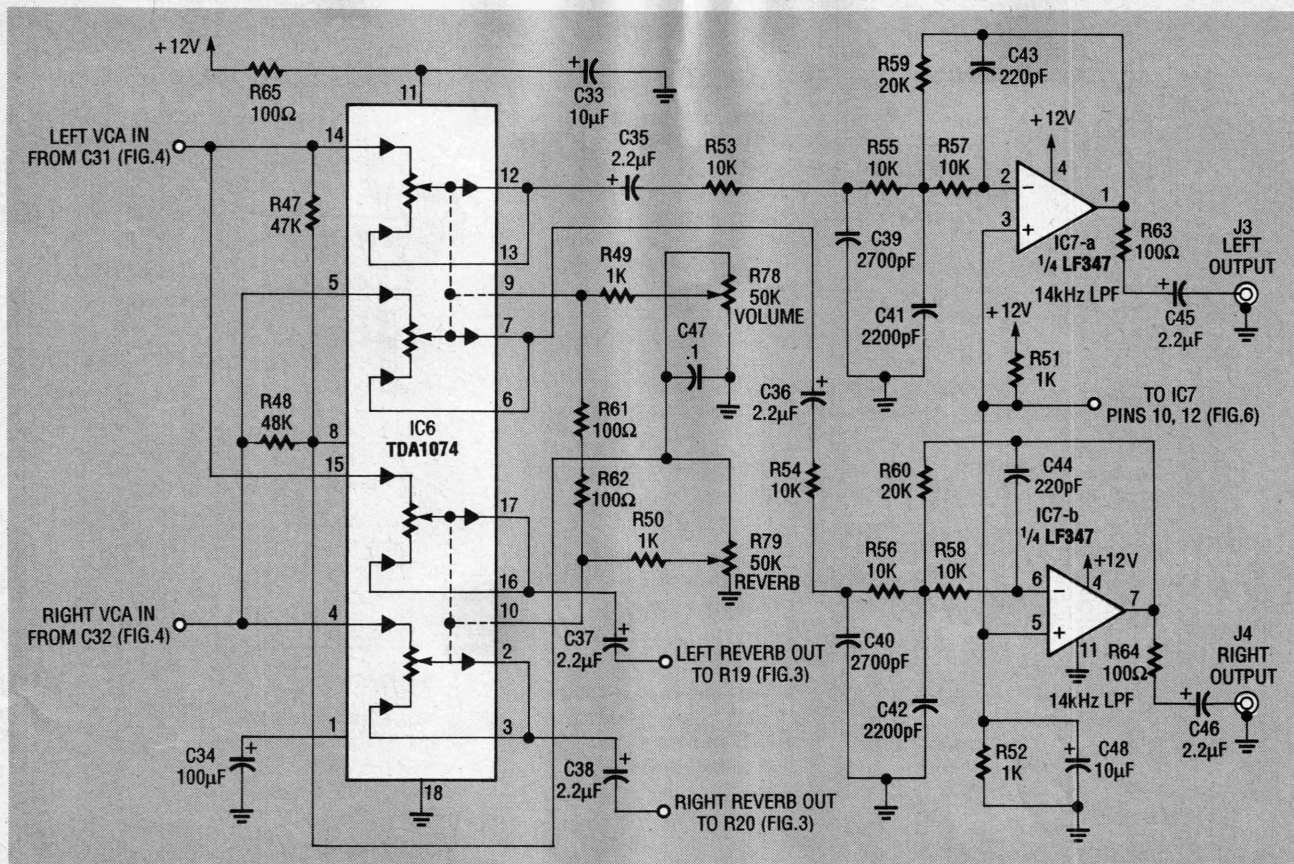
The BBD is rated by the manufacturer to have an upper frequency response limit of 10 kHz and a signal-to-noise ratio of 55 dB at a clock frequency of 40

kHz. (Note that the S/N ratio of a BBD improves as the clock frequency rises.) Actual testing of the device has proven that it will work slightly beyond 12 kHz. In the AFG II, the anti-aliasing low-pass filter (LPF) and post-BBD LPF were designed to obtain the maximum performance available from the BBD.

Returning to the circuit diagram Fig. IC1-b and IC1-c form the anti-aliasing, low-pass BBD input filter. Although basically a second-order filter, component values have been selected to obtain a response curve which rises smoothly from 0 dB at 1 kHz until it peaks at about +5 dB at 10 kHz. The response then falls back to 0 dB at 14 kHz, and continues to drop at about 6 dB per octave. In Fig. 4, IC5-a and IC5-b form the post-BBD low-pass filter. Component values for this third-order Butterworth filter have been selected to produce a response curve nearly opposite that of the input filter circuit. This method of high-frequency

pre-emphasis followed by de-emphasis of the audio signal passing through the BBD results in an improved signal-to-noise ratio for the delay section of about 5 dB. The actual measured frequency response of the entire AFG II from input to output varies no more than 1 dB from 20 Hz to 10 kHz. The -3 dB point is 12 kHz, and response beyond 12 kHz drops at more than 36 dB per octave.

The next section, shown in Fig. 5, is the volume and reverb control stage. A TDA1074 two-section dual-ganged electronic volume control, or EVC (IC6), eliminates the need for an expensive dual-ganged potentiometers, as well as all the wires required to hook them up. Instead, a single control voltage applied to the control input of the EVC simultaneously adjusts the gain of a matched pair of amplifiers. The range of adjustment available is from infinite attenuation to unity gain. The VOLUME control (R78) adjusts the level of the signal being sent to the final output filter stage. The REVERB control (R79) adjusts the level of



the delayed signal being fed back to IC1-b and IC1-c where it is added back to the input signal to produce the reverberation effect. Components R17, R19, and C13, and R18, R20, and C14 form a pair of first-order LPF's to prevent ringing and match levels of the feedback signal. The output stage, IC7-a and IC7-b, is another pair of third-order LPF's with a corner frequency of 14 kHz. They provide for additional reduction of clock noise from the output signal as well as providing a low output-source impedance.

The other half of IC7 (IC7-c and IC7-d) forms an active crossover network for driving a subwoofer system; IC7-c sums the left and right channel inputs, inverts the summed signals by 180°, and provides a low driving impedance for the following low-pass filter stage (see Fig. 6). Potentiometer R76 adjusts the gain of this stage from unity to 3 times unity value. Op-amp IC7-d and its associated

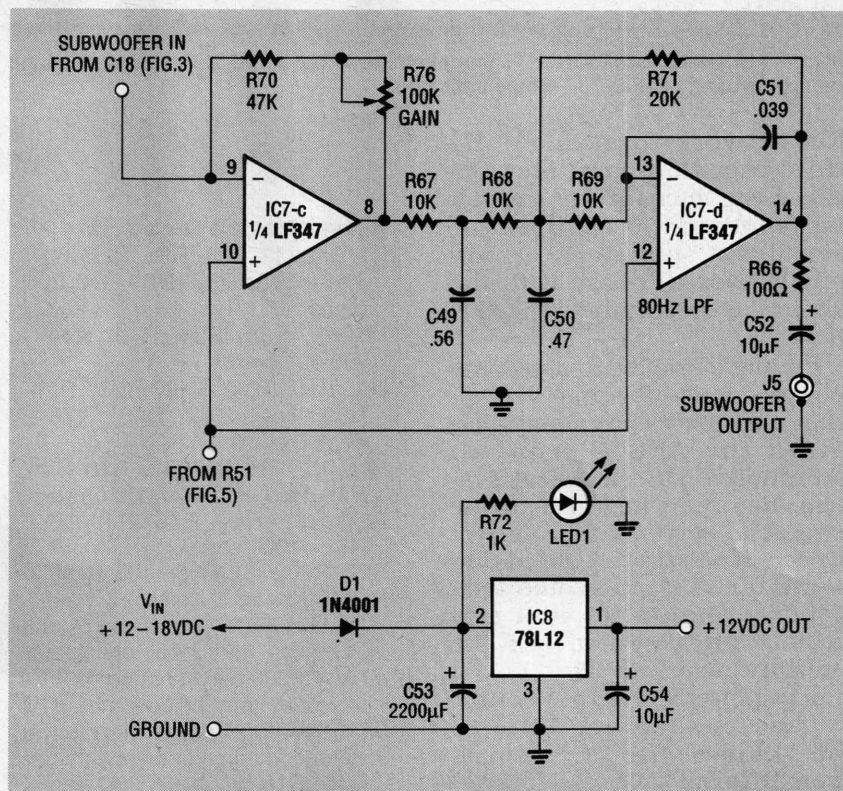


FIG. 6—AN ACTIVE CROSSOVER NETWORK for driving a subwoofer system is formed by IC7-c and -d. Potentiometer R76 adjusts the gain of this stage.

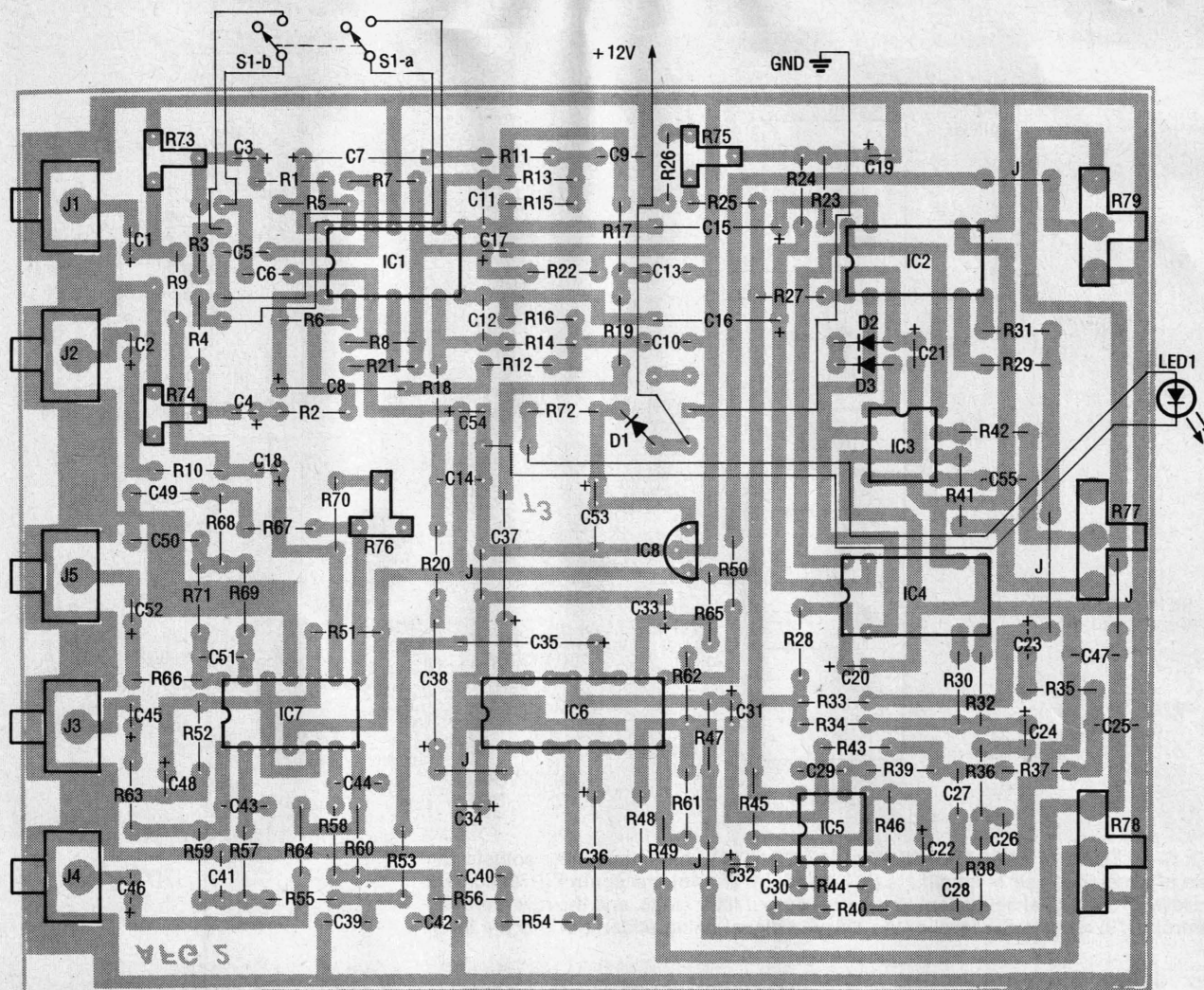


FIG. 7—ABOUT 150 COMPONENTS mount on the AFG II board. Don't forget to install the six jumpers, marked "J," where indicated.

RC network form an 80-Hz third-order low-pass filter. Because the filter inverts the signal another 180 degrees, the signal that appears at the output is back in phase with the signals at the inputs to the AFG II.

Finally, a brief word about the power supply. Rectifier D1, although unnecessary when operating the AFG II from an automotive DC electrical system, has been included in the circuit to serve as a protective device; it will protect about \$50 worth of IC's if you should accidentally connect the unit to a car's electrical system with the polarity reversed. The AFG II can be powered from any source capable of supplying 14 to 18 volts of noise-free DC. A small, well-filtered DC wall transformer is all that's required to bench test the AFG II, or adapt it

PARTS LIST

All resistors are 1/4-watt, 5%.

R1-R4, R13, R14, R25, R26, R41, R53-R58, R67-R69—10,000 ohms
 R5-R8, R33, R34, R43, R44—100,000 ohms
 R9, R10, R17, R18, R23, R24, R35-R40, R47, R48, R70—47,000 ohms
 R11, R12, R15, R16, R19, R20, R42, R59, R60, R71—20,000 ohms
 R21, R22, R45, R46, R49-R52, R72—1000 ohms
 R27, R28, R61-R66—100 ohms
 R29-R32—5600 ohms
 R73, R74—1000 ohms, horizontal PC-mount trimmer potentiometer
 R75—10,000 ohms, horizontal PC-mount trimmer potentiometer
 R76—100,000 ohms, horizontal PC-mount trimmer potentiometer
 R77, R78, R79—50,000 ohms, PC-mount potentiometer

Capacitors

C1-C4, C18, C23, C24, C31, C32, C45, C46—2.2 μ F, 50 volts, radial electrolytic
 C5, C6—4700 pF, 50 volts, Mylar or metal film, 5%
 C7, C8, C15, C16, C35-C38—2.2 μ F, 63 volts, axial electrolytic
 C9, C10, C13, C14—3900 pF, 50 volts, Mylar or metal film, 5%
 C11, C12, C43, C44—220 pF, ceramic disc, 5%
 C17, C19, C20-C22, C33, C48, C52, C54—10 μ F, 35 volts, radial electrolytic
 C25, C26—560 pF, ceramic disc, 5%
 C27, C28—470 pF, ceramic disc
 C29, C30—47 pF, ceramic disc, 5%
 C34—100 μ F, 16 volts, radial electrolytic
 C39, C40—2700 pF, 50 volts, Mylar or metal film, 5%

BUILD THIS

STUN GUN

ROBERT GROSSBLATT, and ROBERT IANNINI

Man's fascination with high voltages began with the first caveman who was terrified by a bolt of lightning. In more recent times, electronics experimenters and hobbyists have found the Tesla coil and the Van de Graaff generator equally fascinating. In this article we'll show you how to build a hand-held high-voltage generator that is capable of producing 75,000 volts at a power level as high as 25,000 watts. The stun gun can be used to demonstrate high-voltage discharge and as a weapon of self-defense. Before building one, however, you should read and pay very close attention to the warning in the accompanying text box, as well as to the description of physiological effects that follows.

This experimental high-voltage generator can produce 75,000 volts at a peak power of 25,000 watts.



WARNING

THIS DEVICE IS NOT A TOY. We present it for educational and experimental purposes only. The circuit develops about 75,000 volts at a maximum peak power of 25,000 watts. The output is pulsed, not continuous, but it can cause a great deal of pain should you become careless and get caught between its output terminals. And you should never, repeat, NEVER, use it on another person! It may not be against the law in your area to carry a stun gun in public, but, if you use it on another person, you may still be liable for civil action.

To help you build, test, and adjust the device safely, we have included a number of tests and checks that must be followed strictly. Do not deviate from our procedure.

Physiological effects

So that you may understand the danger inherent in the stun gun, let's discuss the physiological effects first. When a high voltage is discharged on the surface of the skin, the current produced travels through the nervous system by exciting single cells and the *myelin sheaths* that enclose them. When that current reaches a synapse connected to a muscle, it causes the muscle to contract violently and possibly to go into spasms.

The longer contact with the high voltage is maintained, the more muscles will be affected. If the high voltage maintains contact with the skin long enough to cause muscle spasms, it may take ten or fifteen minutes before the brain is able to re-establish control over the nerve and muscular systems.

How much power is required to cause such spasms? That's not an easy question to answer because, although it is relatively easy to make precise measurements of the power produced by a high-voltage device, it is difficult to rate the human body's susceptibility to shock accurately. Some obvious factors include age and diseases such as epilepsy. But the bottom line is simple: *The only one who fools around with a stun gun is a fool.*

The amount of energy a device delivers is actually the amount of power delivered in a given period of time. For our purposes, it makes sense to talk about energy in joules (watt-seconds). Using a fresh 9.8-volt Ni-Cd battery, the stun gun is capable of delivering peak power pulses of 25,000 watts. Actually, pulses start out at peak power and then decay exponentially. The length of the decay time depends on the components used in the circuit, the ambient temperature, the battery's capacity, and the positioning of the output contacts with respect to each other.

Assuming that the decay rate is purely exponential, the stun gun can produce about 0.5 joules of energy, provided that the battery is fully charged. Let's put that number in perspective.

Both the Underwriter's Laboratory (in Bulletin no. 14) and the U. S. Consumer Product Safety Commission state that ventricular fibrillation (heart attack) can be caused in humans by applying 10 joules of energy. Since the stun gun only generates about half a joule, you might think that a device that produces only one twentieth of the critical amount has a more-than-adequate margin of safety. *Don't bet on it.* A brief contact with the stun-gun's discharge hurts a great deal, but it takes only about five seconds of continuous discharge to immobilize someone completely.

Let's compare the stun gun's output with a similar device, called a Taser gun, which appeared on the market a few years ago. You may have seen a film demonstrating just how effective the Taser could

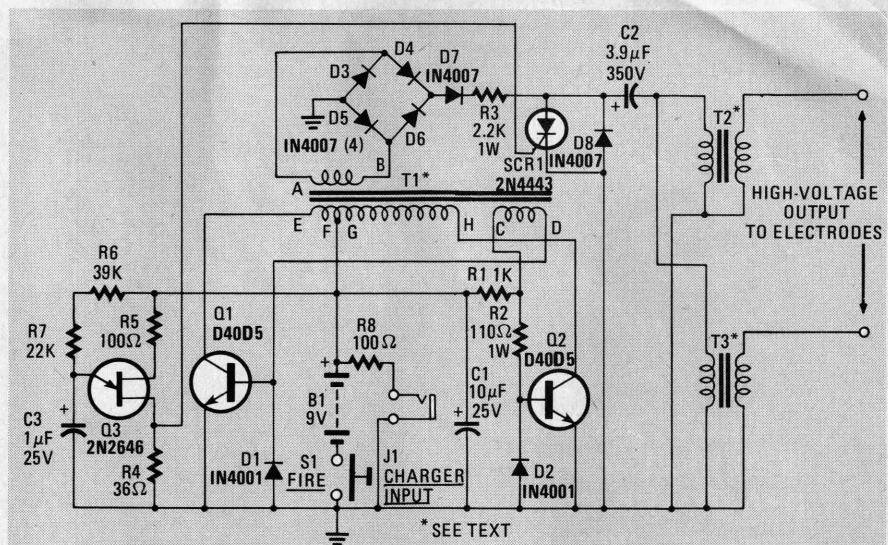


FIG. 1—THE STUN GUN'S CIRCUIT is a multi-stage voltage step-up circuit. The Q1/Q2 circuit produces a squarewave output of about 10 kHz, and Q3 produces 15- μ s discharge pulses at a rate of about 20 ppm. Those pulses fire SCR1, which induces a voltage in the windings of step-up transformers T2 and T3.

be as a deterrent. A foolhardy volunteer was paid an enormous sum of money to have the Taser fired at him. No matter how big, strong, (and stupid) the person was, as soon as the Taser's "darts" hit him, he would collapse to the ground and go into uncontrollable convulsions.

The energy produced by the Taser is only 0.3 joules—about 60% of what our stun gun produces! Even so, the Taser has been officially classified as a firearm by the Bureau of Tobacco and Firearms because it shoots its electrode "darts" through the air. Even though our stun gun doesn't operate that way, the Taser puts out considerably less energy than the stun gun. Keep those facts and figures in mind as you assemble and use the device.

How it works

The schematic diagram of the stun gun is shown in Fig. 1. Basically, it's a multi-stage power supply arranged so that each succeeding stage multiplies the voltage produced by the preceding stage. The final stage of the circuit feeds two oppositely-phased transformers that produce extremely high voltage pulses. If that description sounds familiar, you've probably studied capacitive-discharge ignition systems—the stun gun works on the same principles.

The first section of the power supply is a switcher composed of Q1, Q2, and the primary windings (connected to leads E, F, G, and H) of T1. When FIRE switch S1 is closed, R1 unbalances the circuit and that causes it to start oscillating. Since base current is provided by a separate winding of T1 (connected to leads C and D), the two transistors are driven out of phase with each other, and that keeps the circuit oscillating. Resistor R2 limits base drive to a safe value, and diodes D1 and D2 are steering diodes that switch base current

from one transistor to the other. Oscillation occurs at a frequency of about 10 kHz.

The switching action of the first stage generates an AC voltage in T1's high-voltage secondary (leads A and B). The amount of voltage depends on the battery used, but a battery of seven to nine volts should produce 250 to 300 volts across T1's secondary.

That voltage is rectified by the full-wave bridge composed of diodes D3–D6. Capacitor C2 charges through D7 at a rate that is controlled by R3.

The value of capacitor C2 affects the output of the stun gun. The greater the capacitance, the more energy that can be stored, so the more powerful the discharge will be. A larger capacitor gives bigger sparks, but requires more charging time, and that gives a lower discharge rate. On the other hand, a smaller capacitor gives smaller sparks, but a faster discharge rate. If you wish to experiment with different values for C2, try 3.9 μ F (as shown in Fig. 1), 7.8 μ F, and 1.95 μ F. Those values were arrived at by using one 3.9 μ F capacitor alone, two of the same capacitors in series, and two in parallel.

Meanwhile, UJT Q3 produces 15- μ s pulses at a rate of about 20 ppm. That rate is controlled by C3 and the series combination of R6 and R7. When a pulse arrives at the gate of SCR1, it fires and discharges C2. That induces a high-voltage pulse in the primary windings of T2 and T3, whose primaries must be wired out of phase with each other. The result is a ringing wave of AC whose negative component then reaches around and forces the SCR to turn off. When the next pulse from Q3 arrives, the cycle repeats.

The outputs of the stun gun appear across the secondaries of T2 and T3. The hot leads of those transformers connect to

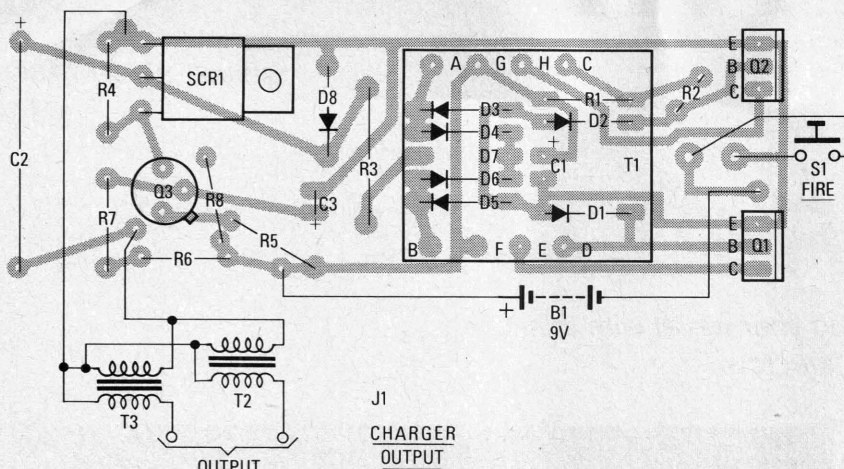


FIG. 2—MOUNT ALL COMPONENTS ON THE PC BOARD as shown here. Note that T2 and T3 are mounted off board, and that J1, C1, and D7 mount on the foil side of the board. In addition, a number of components mount beneath T1: D1-D6, R1, and R3. Those diodes and resistors must be installed before T1.

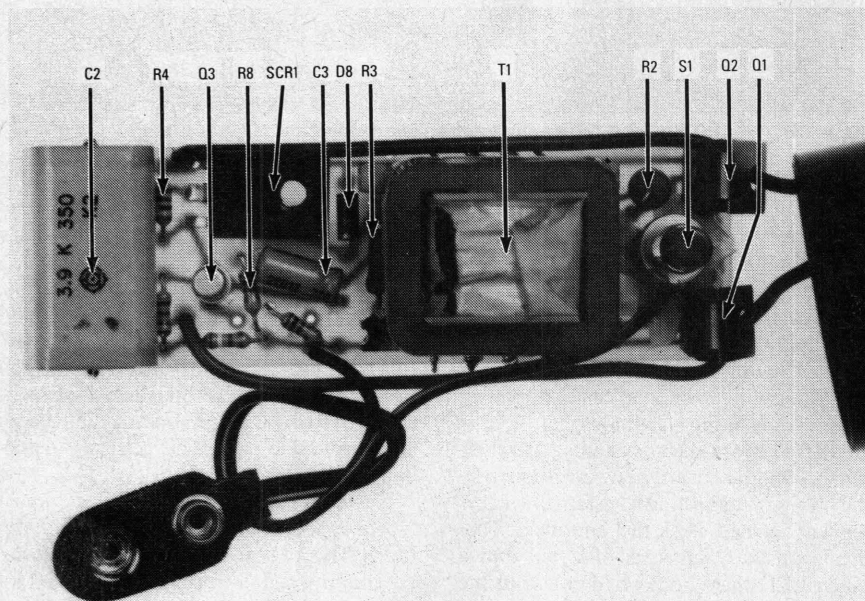


FIG. 3—BEND SCR1'S LEADS 90° so that the nomenclature faces up and then solder the SCR to the board. Also note that C3 must be bent over at a 90° angle, and that R2 is mounted vertically.

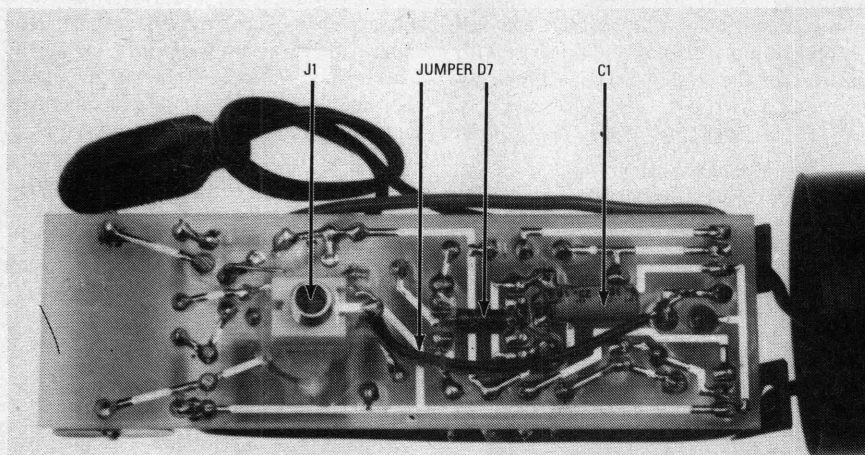


FIG. 4—JACK J1, DIODE D7, AND CAPACITOR C1 mount on the foil side of the PC board. One terminal of J1 mounts to the same pad as R8, and the jack should be glued to the board with RTV (or other high-voltage compound) after you verify that the circuit works properly.

PARTS LIST

All resistors are 1/4-watt, 5% unless otherwise noted.

R1—1000 ohms
R2—110 ohms, 1 watt
R3—2200 ohms, 1 watt
R4—36 ohms
R5, R8—100 ohms
R6—39,000 ohms
R7—22,000 ohms

Capacitors

C1—10 μF, 25 volts, electrolytic
C2—3.9 μF, 350 volts, electrolytic
C3—1 μF, 25 volts, electrolytic

Semiconductors

D1, D2—1N4001, 50-volt rectifier
D3-D8—1N4007, 1000-volt rectifier
Q1, Q2—D40D5, power transistor
Q3—2N2646, UJT
SCR1—2N4443

Other components

B1—9-volt Ni-Cd battery
S1—SPST momentary pushbutton switch
T1—12 to 400 volts saturable-core transformer. See text
T2, T3—50 kilovolt pulse transformer, 0.32 joules. 400-volt primary. See text

Note: The following components are available from Information Unlimited, P. O. Box 716, Amherst, NH 03031: T1, \$12.50; both T2 and T3, \$12.50; C2, \$1.50; PC board, \$4.50; case, \$3.50; case with T2 and T3 potted, \$17.50; charger, \$6.50; 9.8-volt battery, \$16.50; complete kit of all parts including all components, PC board, case, and charger, but no battery, \$39.50.

the output electrodes, which should be held securely in position about two inches apart, and which should be insulated from each other and from the environment with high-voltage potting compound.

Batteries

The stun gun can be powered with almost any battery that can supply at least seven volts at one amp. A Ni-Cd battery would be a good choice; R8 and J1 will allow the battery to be recharged without removing it from the case.

The higher the battery's voltage, the higher the stun-gun's output voltage. Most nine-volt Ni-Cd's actually have a maximum fully-charged output of only 7.2 volts. However, batteries that deliver 9.8 volts when fully charged are available from several sources.

Construction

Keep in mind the fact that the stun gun produces dangerously high voltages, and don't approach the construction of the stun gun with the same nonchalance with which you might build a light dimmer.

The circuit can be built on a PC board or on perfboard. The foil pattern for a PC

continued on page 94

STUNGUN

continued from page 43

board is shown in "PC Service;" alternatively, a PC board can be purchased from the source mentioned in the Parts List. If you build the circuit on a per-board, follow our parts layout closely; otherwise you may have problems with arcing.

Due to the critical nature of the three transformers, we are not providing details on winding them. They are available from the source mentioned in the Parts List.

Referring to the parts-placement diagram in Fig. 2, and the photos in Fig. 3 and Fig. 4, mount all components except C2, T1, T2, and T3 on your board. Note that several components mount on the foil side of the PC board: C1, D7, and J1. Do not install those parts yet either.

After all components (except those mentioned above) are installed, check your work very carefully, especially D1-D6, R1, and R3, because T1 will be installed above them, and there will be no chance to correct errors later. After you're absolutely sure that they're installed correctly, install T1 with the black mark on the windings mounted toward C2.

Foil-side components

One of J1's tabs shares a hole on the PC board with resistor R8, which should be

mounted already. Solder the tab of J1 that corresponds to the tip (not the barrel) of an inserted plug to the indicated pad. Then mount C1 and D7. Last, solder a 1 $\frac{3}{4}$ -inch piece of 18-gauge wire to the barrel pin of J1, and connect the opposite end of that wire to the appropriate pad beneath S1, the FIRE SWITCH.

Preliminary check-out

WARNING: While measuring voltages and currents, keep your face, hands, and all metallic objects away from the high-voltage end of the stun gun. If you want to prod a component, use a non-conductive rod such as a plastic TV alignment tool. High voltage behaves very differently than low voltage. Any material that retains moisture can serve as a discharge path. THAT INCLUDES WOOD! Also, never work on or use the unit when your hands are wet.

Connect a voltmeter (set to a 1000-volt DC range) to ground and to the output of the D3-D6 diode bridge. Then power up the circuit using either a freshly-charged battery or an external supply capable of delivering 9.8 volts at one amp. If everything is working properly, you should measure about 400-volts DC at the output of the bridge when you press S1.

If you don't measure that voltage, connect an oscilloscope to the collector of Q1 or Q2. You should see a squarewave with a

period of about 100 μ s. If that waveform is not present, the switching circuit is not operating correctly. *Remove power and check your wiring again. Do not debug the circuit with a battery connected!*

Resistor R6 controls the rate at which the UJT (Q3) discharges, and R3 controls the rate at which C2 charges. You can experiment with the values of those components if you are not satisfied with the circuit's high-voltage output. R3 can vary from 2.2 to 4.7K. You can also experiment with the value of C2. See Table 1.

After the circuit is operating correctly, attach J1 to the board with high-voltage potting compound or RTV. And before you mount the circuit in a case, make sure there's no arcing on the PC board. If there is, you can stop it with a liberal application of RTV, paraffin, or epoxy.

Conclusion

The stun gun's discharge is very impressive. The spark is highly visible and each discharge produces a sharp, resounding crack. The circuit can teach you much about voltage-multiplying circuits and power supply design. But don't ever forget that the stun gun is not a toy. It can cause much damage to both you and others. Never leave it lying around where children, pets, or anyone unfamiliar with how to use it can handle it. It's a good idea to remove the battery before storing the stun gun. Above all: be careful! **R-E**

BRAINSTORM

continued from page 62

no additional information, as it is indicated in the schematic; instead, it would only clutter up the chassis drawing.

Circuit board and schematic

Many projects begin with a vague idea for a device followed by some rough circuit sketches. That's OK for the diddle stage, but when its time to generate the schematic, the diagram that will control the building stage of the project, the drawing must be exact and complete.

Once you've finished the schematic, you need to design the circuit-board layout. As you're no doubt aware, when you draw a schematic, the symbols used bear no relationship to the actual size of the components themselves. Thus, while a resistor and capacitor may appear to occupy areas that are roughly equal, in reality the resistor may be only a fraction of the size of the capacitor, especially if the capacitor is a large electrolytic. To be sure that you've allowed sufficient space for each component in your design, use the actual parts and lay them out on an actual-size drawing of the circuit board. That is especially important when design-

ing PC-board layouts. See "Designing Double-Sided Printed Circuit Boards," in the September 1985 issue of **Radio-Electronics** for tips on laying out complicated circuitry. The circuit-board drawing should include identification of the connection points to any off-board components.

Procedure

You've prepared the paperwork, and you've assembled a kit with all of the parts. Now, you're ready to put the circuit together, turn it on, and watch for smoke. When the building process begins, you switch from designer to technician, with your paperwork guiding you every step of the way.

The following step-by-step procedure applies the paperwork to the construction job, and covers initial assembly to finished product. Steps 1 and 2 cover breadboarding individual circuits for design debugging, and will be repeated until each circuit works on the experimenter's solderless breadboard. Once the circuit operates correctly, final assembly requires repeating steps 1 and 2 to rebuild the circuit in its final form.

Step 1—Mount the components on your circuit board (solderless experimenter, perforated, wire wrap, or etched) using the circuit-board drawing as a

guide.

Step 2—Do the wiring. When using wire-wrap or point-to-point techniques, as each connection is made, trace over the appropriate line on the schematic using a colored pencil.

For PC boards, the same technique should be followed, but it should be done while you are designing the board. As a trace is laid down, the line or lines on the schematic should be traced over.

Completely test the board using a temporary rig to mount any off-board parts. When the circuit passes all of your tests, it's ready for installation.

Step 3—Install the panel-mounted parts.

Step 4—Install the circuit board or boards.

Step 5—Wire the chassis.

Keep the wiring as short and as neat as possible. Use wiring ties, cable clamps, etc.

Step 6—Apply power. If you've been very careful, and followed the steps we've shown you, the odds of getting a correctly working circuit the first time are greatly improved. Of course, they are not.

Fortunately, if you've done your paperwork properly, you will have a paper trail to follow if you run into trouble. Very often that trail will lead you directly to the cause of your problem. **R-E**

SINGLE TRANSISTOR SWITCHING CIRCUIT DESIGN

Jeff Holtzman

■ Hobbyists dealing with digital circuits are terrified by interfacing. But for computers to do useful things, they must have a way of talking with outside circuits: relays, motors, lightbulbs, LED's, etc. Things aren't so bad when dealing with standard TTL outputs, but what do you do when you have to connect a 5-horsepower motor to a CMOS inverter? Doing that is simple; we'll show you how to do the design calculations, and we'll present a short BASIC program that does the work. One feature of that program is that after making the calculations, it outputs the value of the closest standard resistor.

To design a switching circuit, before doing any calculations, make sure that a single transistor will do the job. There are no calculations involved, just a little common sense. If your driving device is a 6522 VIA, or a Z80 PIO, or similar, and you want to control an NC (Numerical Control) lathe with a 5-horsepower motor, you'll need a few transistors to build up enough drive to control a relay large enough.

If you need to turn an LED on and off, you may be able to use a transistorless driving circuit. You may get by using just a current-limiting resistor, although doing that may not allow the LED to glow at full brightness.

If a one-transistor circuit will do the job for you, choose a transistor. Double the voltage and current requirements of your load device, and then search your databooks and catalogs for a transistor that meets those specifications. Find several that will do, in case your first choice is unavailable, or expensive.

Every transistor has a DC-gain factor, called HFE. Unfortunately, HFE varies from device to device, even between devices of the same type. HFE also varies with load and temperature. Since many applications aren't critical, a few assumptions can be made to get

things off the ground. Those will allow you to arrive at tentative values suitable for breadboarding; remember that slight adjustments may have to be made later.

With specifications handy, choose a value for HFE halfway between the minimum and maximum values given in the databook. If you don't have the specs, you can assume a value of 100 for a small-signal transistor, or 10 for a power transistor. That lets you estimate the value of the current-limiting resistor in the base circuit (R1 in Fig.1). As base current is defined as the collector current over HFE, you can get the base current by dividing the collector current by 100, or by 10, depending on the type of transistor. Before finding the value of R1, we have to find the value of R2.

Making the calculations

To calculate the values of the resistors in the switching circuit shown in Fig. 1, work backwards from the load. You have to know how much current the load will draw, and its resistance, to calculate the voltage drop, V(L), that will appear across it. According to Kirchoff's law,

$$V(CC) = V(L) + V(R2) + V(CE)$$

V(CE) is the voltage dropped across the collector and emitter leads of Q1; that voltage is often about 0.3 volt, so it may be ignored. By making that simplification and rearranging our equation we find that $V(R2) = V(CC) - V(L)$. We need the resistance of R2, and by Ohm's law that is equal to the voltage across R2 divided by the current flowing through it, or $R2 = (V(CC) - V(L)) / I(L)$. If the load is an LED, we can assume that V(L) is 1.5 volt. If not, V(L) must be expressed as $I(L) * R(L)$. In the latter case, the value of R2 would be expressed as $R2 = (V(CC) - [I(L) * R(L)]) / I(L)$.

To calculate the value of R1, we use the value of HFE we found above. Since $HFE = I(C) / I(B)$, I(B) can be expressed as $I(B) = I(C) / HFE$.

By Thevenin's law we know that I(C) must equal I(L), so $I(B) = I(L) / HFE$. Again, by Kirchoff's law we know that $V(IN) = V(R1) + V(BE)$, where V(BE) is the base-emitter voltage drop. For the sake of simplicity, we can assume that V(BE) is zero, so the voltage dropped across R1 must equal the input voltage. Therefore, the value of R1 can be simply expressed as $R1 = V(IN) / I(B)$. Since I(B) can be expressed in terms of collector current, $R1 = V(IN) / [I(C)/HFE]$.

The unknown quantities (R1 and R2) are expressed in terms of the known quantities, so it's time to plug the values into the equations and solve them. Rather than bang those values out your calculator, type the program shown in Listing 1 into your computer and save it for the next time you need to design a switching circuit.

The program provides features you may find convenient. For example, that program can be used to design a circuit with any driving voltage, but if you happen to be using TTL, the program assumes a value of 2.4 volts for V(IN). Likewise, if you will be driving an LED, the program assumes a load current of 15 mA, and a 1.5 volt drop across the LED.

The program was written in MBASIC-80, and it should be simple to translate into other dialects of BASIC. The main routines have been written as


```

10 REM switching transistor circuit
20 REM design jh 10-22-85 for RE
30 FIRST = -1
40 PRINT CHR$(27);"*";
50 GOSUB 1000
60 PRINT
70 INPUT "Another? ", YESNO$
80 IF LEFT$(YESNO$,1) = "Y"
   OR LEFT$(YESNO$,1) = "y"
   THEN GOTO 50
90 END
1000 REM get and print data for
    switching circuit
1010 PRINT
1020 INPUT "Enter supply voltage
(VCC) in volts: ", VCC
1030 INPUT "Enter HFE of
transistor: ", HFE
1040 INPUT "Is load an LED
(y/n) ? ", YESNO$
1050 YESNO$=LEFT$(YESNO$,1)
1060 IF YESNO$="Y"
   OR YESNO$="y"
   THEN VL=1.5: IL=15: GOTO 1100
1070 INPUT "Enter load
resistance (RL): ", RL
1080 INPUT "Enter load
current (IL) in mA: ", IL
1090 VL=IL*.001*RL
1100 IL=IL*.001
1110 INPUT "Does TTL drive this
circuit (y/n) ? ", YESNO$
1120 YESNO$=LEFT$(YESNO$,1)
1130 IF YESNO$="Y"
   OR YESNO$="y"
   THEN VIN = 2.4: GOTO 1150
1140 INPUT "Enter driving
voltage (VIN): ", VIN
1150 IB=IL/HFE
1160 R1=VIN/IB
1170 RIN=R1:GOSUB 10000:R1=ROUT
1180 R2=(VCC-VL)/IL
1190 RIN=R2:GOSUB 10000:R2=ROUT
1200 PRINT:PRINT "R1 =";R1;
    "ohms, R2 =";R2;
    "ohms, I(B) =";IB*1000;"ma"
1210 RETURN
10000 REM get closest standard
    resistor value
10010 REM input = rin
10020 REM output = rout
10030 REM sf = scale factor
10040 REM 25 standard values follow
10050 DATA 1.0, 1.1, 1.2, 1.3,
    1.5, 1.6, 1.8
10060 DATA 2.0, 2.2, 2.4, 2.7
10070 DATA 3.0, 3.3, 3.6, 3.9
10080 DATA 4.3, 4.7
10090 DATA 5.1, 5.6
10100 DATA 6.2, 6.8
10110 DATA 7.5, 8.2, 9.1, 10.0
10120 IF NOT FIRST THEN 10190
10130 FIRST = 0
10140 AMAX = 25 : REM must equal
    no of entries in table above
10150 DIM A (AMAX)
10160 FOR I = 1 TO AMAX
10170 READ A(I)
10180 NEXT I
10190 X = RIN
10200 SF = 0
10210 WHILE X > A(AMAX)
10220 X = X/10
10230 SF = SF + 1
10240 WEND
10250 I = 1
10260 WHILE X > A(I)
10270 I = I + 1
10280 WEND
10290 IF X-A(I-1) <= A(I)-X
    THEN X = A(I-1) ELSE X = A(I)
10300 ROUT = X * 10^SF
10310 RETURN
10320 END

```

subroutines for easy integration into your programs. Lines 1000-1210 get the data and calculate the resistor values; lines 10000-10320 select the closest standard resistor values. Lines 10-90 drive the main routine following line 1000. Line 30 sets a flag for the standard-value routine, so that it doesn't read the data contained in lines 10050-10110 in every time a set of calculations are made. Line 40 clears the screen of my CRT (an ADM 20); you'll probably have to use another method.

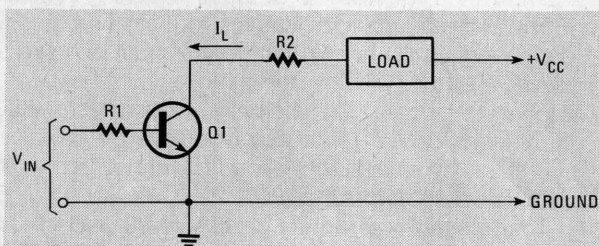


FIG. 1—R1 IN DIAGRAM ABOVE is used for current limiting and must be estimated. (See text.)

If your BASIC doesn't have the WHILE . . . WEND statements, you can simulate them by using an IF . . . THEN statement and a dummy loop. For example, the code below could be substituted for the loop running from line 10210 to line 10240: 10210/IF X <= A(MAX) then 10250

```

10220 X = X/10
10230 SF = SF + 1
10240 IF X > A(AMAX) THEN 10210

```

That's about all there is to calculating the values of the resistors in this simple switching circuit. Of course, you should verify that the calculated values actually do the job before building 10,000 duplicate circuits. Also, note that the program prints the value of I(B) the circuit will draw. That was done as a protective measure. If your driving circuit is CMOS, and I(B) is greater than a milliamp or so, you'll have to do one of two things: use a Darlington, or add another transistor that will provide a bit of "preamplification." If you opt for a Darlington, you can use our program as is. ◀▶

BUILD THIS

LOW-BAND CONVERTER

There's a world of interesting activity on the frequencies below the AM broadcast-band. Here's a description of what happens down there, and instructions for building a converter for your receiver so you can listen in.



STAN GIBILISCO

THOSE OF US WHO HAVE SHORTWAVE communications receivers have heard, and are familiar with, the range of the radio spectrum between 535 kHz and 30 MHz; that is the extent of coverage of most of today's shortwave receivers. Some receivers have a "longwave" band with a low-frequency limit of around 200 kHz. There aren't many receivers that operate lower than that; the ones that do are rather expensive. This article is concerned with the relatively-little-known part of the electromagnetic spectrum below the standard AM broadcast-band. In particular, we will be looking at the VLF (Very-Low-Frequency) and LF (Low-Frequency) bands, ranging from 3 to 300 kHz. (The VLF band extends from 3 to 30 kHz, and the LF band from 30 to 300 kHz.)

It is not difficult or expensive to receive signals in those frequency ranges. A simple converter can be built from easy-to-find parts for a moderate price, allowing VLF and LF reception with a shortwave communications receiver.

What's below the broadcast band?

There is plenty of activity below 535 kHz, all the way down to 10 kHz in the VLF spectrum. Table 1 shows the frequency allocations below 535 kHz on a worldwide basis. Below 10 kHz, there are no allocations—those frequencies are considered essentially useless, for reasons we will discuss later.

Especially toward the lower part of the LF band (below 150 kHz), and throughout the VLF band, voice-modulated signals will not be found. Such transmissions require too much bandwidth to be used in a part of the spectrum where band-

TABLE 1

Frequency—kHz	Service
Below 10.00	Not allocated
10.00–14.00	Radio location, radio navigation
14.00–19.95	Fixed, maritime mobile
19.95–20.05	Standard frequency
20.05–70.00	Fixed, maritime mobile
70.00–90.00	Fixed, maritime mobile, maritime radio navigation, radio location
90.00–110.0	Fixed, radio navigation, maritime mobile
110.0–130.0	Fixed, maritime mobile, maritime radio navigation, radio location
130.0–160.0	Fixed, maritime mobile
155.0–281.0	Broadcasting (Europe, N. Africa, and Middle East)
160.0–200.0	Fixed
200.0–285.0	Aeronautical radio navigation, aeronautical mobile
285.0–325.0	Maritime radio navigation, aeronautical radio navigation
325.0–405.0	Aeronautical radio navigation, aeronautical mobile
405.0–415.0	Maritime radio navigation, aeronautical radio navigation, aeronautical mobile
415.0–490.0	Maritime mobile
490.0–510.0	Mobile (distress and calling)
510.0–525.0	Mobile, aeronautical radio navigation
525.0–535.0	Mobile, broadcasting, aeronautical radio navigation

width conservation is extremely important. An AM signal takes up at least 6 kHz, and the whole VLF band is only 27 kHz wide! An AM signal at 15 kHz would have to be at least 40 percent as wide as its carrier frequency! Because of those factors, all signals in that frequency range are modulated by means of narrow-bandwidth techniques, such as CW (Morse code) or frequency-shift teletype.

Propagation below 535 kHz

Radio signals at VLF and LF travel by three basic long-distance modes: surface-wave, sky-wave, and waveguide propagation. Particularly below about 100 kHz, the characteristics of radio-signal travel are alien to the short-wave listener; there is no rapid fading, backscatter, or selective distortion such as commonly occurs at high frequencies.

Surface waves

At VLF and LF, radio signals can travel along the surface of the Earth for great distances, without relying on the ionosphere for propagation. This mode of propagation (sometimes mistakenly referred to as 'ground-wave' propagation) gets better and better as the frequency decreases. At 535 kHz, surface waves can be heard out to distances of 200 to 300 miles when conditions are good. But, since the Earth is a poor conductor at that frequency, and the return circuit for surface-wave travel is the Earth, the useful range is limited. Most of the energy gets used to heat up the ground.

As the frequency decreases, the ground becomes a better and better conductor. At frequencies around 100 kHz, it is not unusual to hear surface-wave signals from

more than a thousand miles away. In the VLF range, surface propagation combines with ionospheric propagation to allow worldwide communications, provided that huge antennas and high-power transmitters are used.

Since the ionosphere plays no role in surface-wave propagation, VLF and LF communications may someday prove useful on planets with no ionosphere to support other modes of over-the-horizon links.

Sky waves

All radio waves having frequencies below 535 kHz are dramatically affected by the earth's upper atmosphere. There are three layers of ionized gases high above the surface of our planet; those regions are called the D, E, and F layers. Figure 1 shows the arrangement of those layers during the day and at night. Typical altitudes in miles are shown.

The D layer, at a height of 37 to 57 miles, returns VLF signals to the Earth during the daytime, but absorbs energy at higher frequencies. At night, that layer disappears, and the E and F layers are responsible for VLF and LF sky-wave propagation. The E layer varies in altitude from about 62 to 71 miles and the F layer may be anywhere between about 130 and 261 miles up. The F layer is generally higher at night than during the day.

VLF and LF energy is almost totally reflected by the ionosphere, and hardly any of it escapes into space; furthermore, no VLF or LF signals from outer space can penetrate to the Earth's surface. That creates a "trap" for such energy, and insures that over-the-horizon communication is always possible.

Sky-wave and surface-wave modes, acting together, don't always reinforce each other. If, at a certain distance from the transmitting station, the surface-wave and sky-wave signals are equally strong but opposite in phase, they will cancel each other, and no signal will be heard. That effect becomes more common as the frequency increases.

Also, as frequency increases, the D layer gets more and more absorptive. That reduces the effectiveness of sky-wave communications during the daylight hours. But the D layer disappears at night, and that is why we usually hear LF stations from farther away at night. The same effect occurs throughout the AM broadcast band, and no doubt you have observed the difference between daytime and nighttime propagation there.

Waveguide effect

The ionospheric D layer and the Earth's surface both form almost perfect reflectors of VLF energy. At VLF, the waves are so long that the distance between the ground and the D layer is only a few wavelengths. (For example, a 10-kHz signal has a wavelength of 18.6 miles, but the D layer is 37 to 57 miles

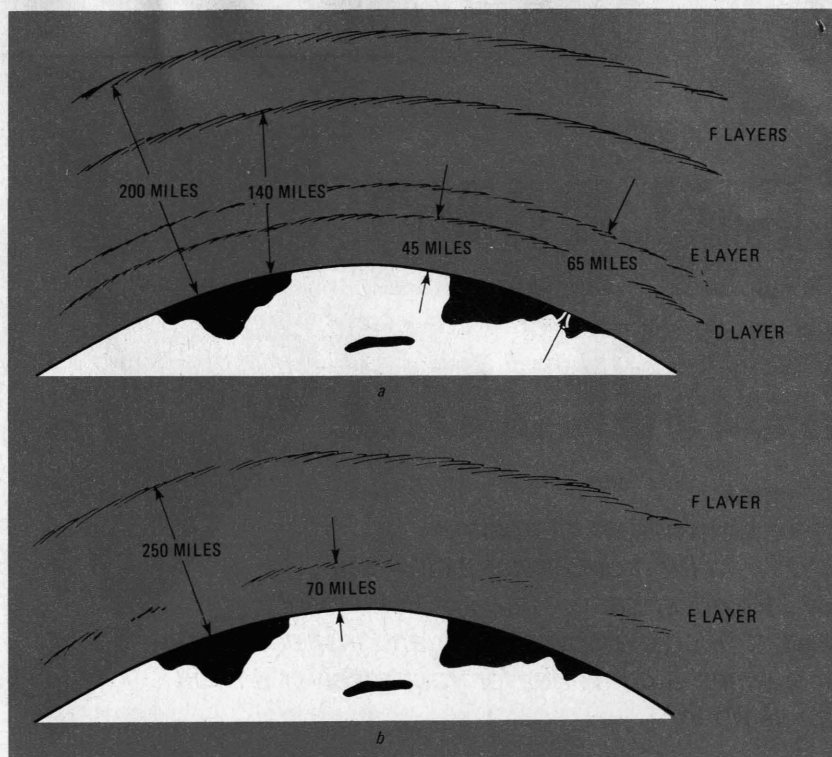


FIG. 1—"D," "E," and "F" LAYERS of the ionosphere assist in LF and VLF propagation. Typical daytime configuration is shown in a; "D" layer disappears at night (b).

high—just two or three wavelengths.) That gives rise to a daytime condition where VLF energy travels within the chamber between the Earth and the D layer as if that space were a waveguide transmission-line.

Waveguide propagation is an extremely reliable means of communication at VLF, with no fading or "dead zones." Waveguide propagation is best during daylight hours, since the D layer disappears at night, leaving only the E and F layers, which are too far above the Earth to serve as good waveguide reflectors.

Since all waveguides have a high-pass frequency response, there is a lower limit on the frequencies that can be used for long-distance propagation. If the wavelength is too great, the Earth-to-D-layer waveguide will be too small to support propagation. The frequency at which attenuation begins to increase rapidly is about 10 kHz. Below that, especially during the daylight hours, the waveguide effect is of little use for long-distance communications. So severe is the loss, in fact, that frequencies below 10 kHz are not allocated for communications.

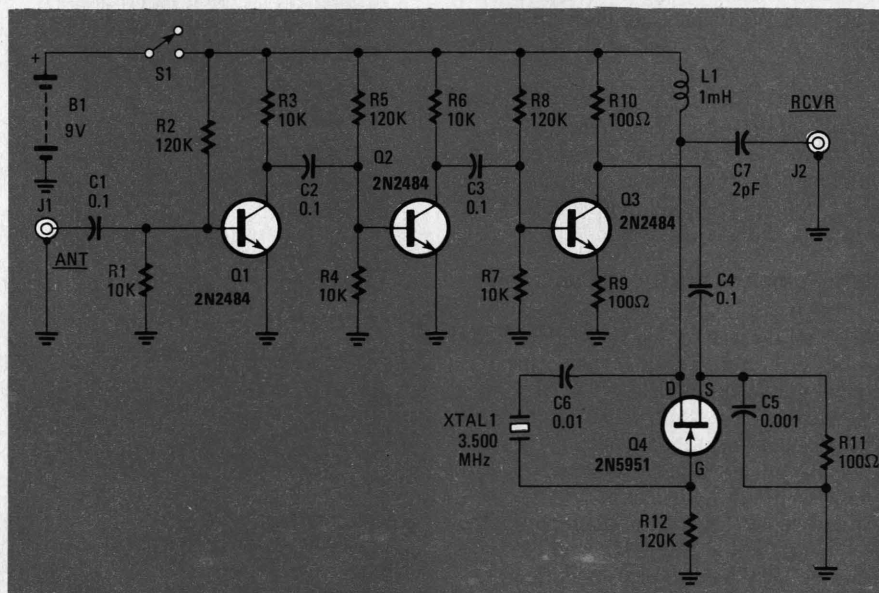


FIG. 2—COMPONENTS USED IN LOW-BAND CONVERTER are readily available. Crystal may be ordered from companies advertising at back of magazine.

Methods of reception

Building a complete, self-contained, receiver that covers 10 to 535 kHz is not easy. The top frequency is more than fifty times greater than the low one; that's equivalent (in terms of percent) to the range from the middle of the AM broadcast-band to TV channel 2! The only practical way in which you can obtain reception over such a wide band is to build a converter for an existing high-frequency receiver.

The converter described here "moves" the 10-to-535-kHz band to the 3.510-to-4.035-MHz range for reception on shortwave receivers. That output range was selected so as to be compatible with amateur-band equipment. The converter's output falls nicely into the 80-meter ham band (3.5-4.0 MHz).

What kind of receiver should you use? It must have a BFO (Beat-Frequency Oscillator). It should be frequency-stable (as drift-free as possible) and have good selectivity. A ham receiver is ideal, especially if it has a narrow-bandwidth filter for CW reception. It also helps to have a noise blanker or limiter, because man-made impulse-noise is a problem at VLF and LF. Other than those requirements, any sort of receiver with fairly accurate dial-calibration is alright: solid-state or tube-type, battery-powered portable or a 50-pound boat anchor from the pre-World-War-II era!

A simple converter

Figure 2 shows the a schematic diagram for a low-band converter. Three stages of broadband amplification (Q1, Q2, and Q3) are used. The crystal oscillator/mixer, Q4, provides an amplitude-modulated (AM) signal, with the carrier at 3.500 MHz. That signal is fed to the antenna input of the receiver. The VLF and LF signals appear as sidebands above and below the carrier. The upper sideband (3.510 to 4.035 MHz) is easier to use because the signal frequency is easier to determine—just subtract 3.500 MHz from the receiver dial read-out. (If you use the lower sideband the reception will be just as good, but the band will come out "upside down" in the receiver, and frequency determination will be tricky.)

The parts for the converter are inexpensive and easy to find. Most, if not all, should be available from advertisers in **Radio-Electronics**. Parts for antenna construction will be described in the text, and are not included in the Parts List.

Construction

The converter circuit can be built on perforated construction board or on an "experimenter board" such as those available from Radio Shack and others; I used the latter.

If you use an "experimenter board," note that the holes are interconnected by foil on the underside of the board in

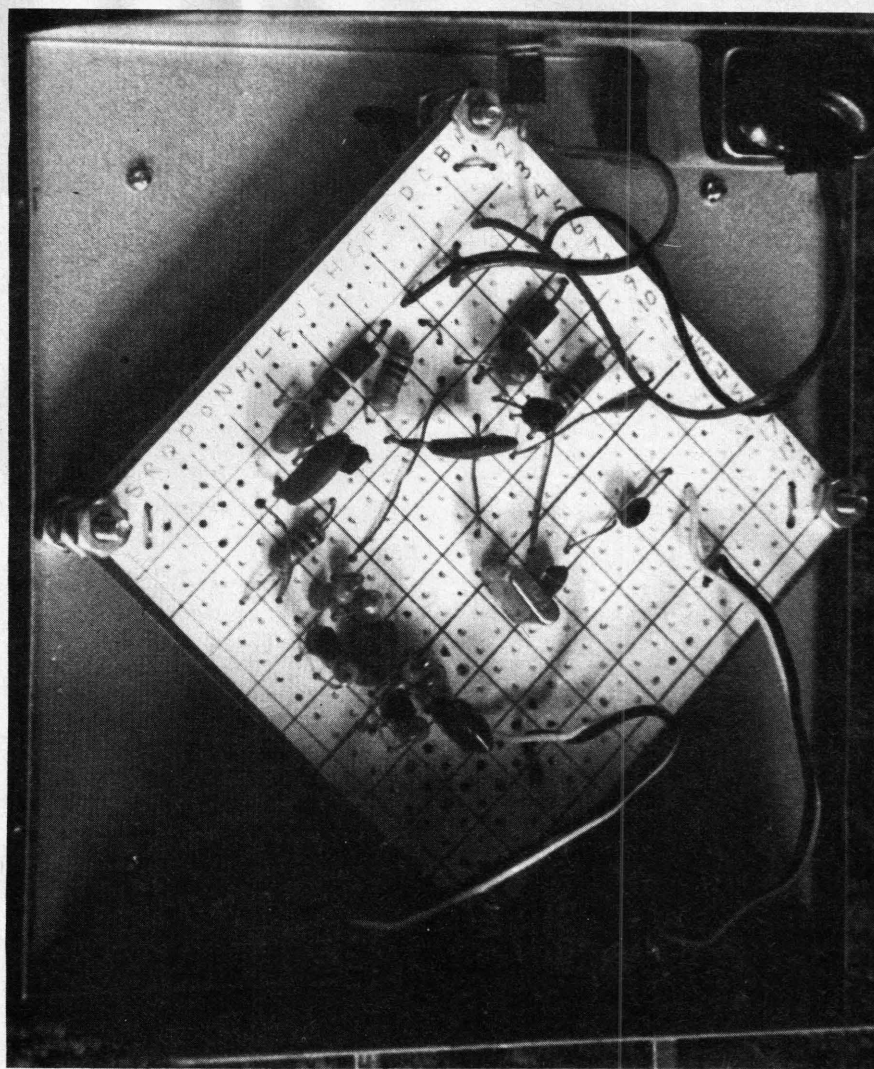


FIG. 3—COMPONENT PLACEMENT is not critical. Letters and numbers at edge of board simplify design work.

groups of four. The 20×20 matrix of holes is labeled on each axis by the numbers 1 through 20 and the letters A through T printed on the underside of the board. It helps to place a piece of adhesive paper on top of the board, punch out all the holes with a sharp instrument, and label the rows and columns for easy reference.

The parts layout I used is shown in Fig. 3. Install the jumper wires first. Then install the resistors, capacitors, crystal, and choke. Resistors should be mounted vertically when the holes are adjacent (as is the case with R1 through R6). Install the transistors and external-connection wires last. Be careful not to use too much heat when soldering the transistor leads.

Drill two $\frac{1}{4}$ -inch holes in the rear panel of the metal cabinet, and one $\frac{1}{4}$ -inch hole in the front panel. Also drill four $\frac{1}{8}$ -inch holes at the corners of a $3\frac{1}{4}$ -inch square (assuming you're using the same type of board I did) on the bottom of the chassis. (It's not critical how the square is oriented, but the holes should be well away from the rubber feet on the underside of the chassis.) Mount the SPST switch on the front panel and the two

female phono-jacks on the rear panel. Don't forget solder lugs for the ground connections to the phono jacks.

The four corner holes on the circuit board must carefully be enlarged with a drill to $\frac{1}{8}$ -inch before mounting. After that's done, put four 6-32 screws into the chassis mounting-holes from the outside of the enclosure, and secure them with one nut apiece. Put a second nut on each screw and move it down until it is $\frac{3}{8}$ -inch from the end.

Push the circuit board down into the screws and, once the board has been pushed down to the middle nuts, screw on the last four nuts to keep it there.

Connect the SPST switch and the input and output jacks to the board; label the jacks to avoid possible confusion later. Connect the 9-volt battery, and secure it in a convenient place with a battery clip or double-sided tape. Be sure none of the wires short to any other circuit point.

Preliminary testing

Now you're ready to test the converter. Use a shielded cable to connect the output of the converter to the antenna terminals (or jack) of the shortwave receiver. Tune

the receiver to 3.500 MHz. When you switch the converter on you should hear a strong unmodulated carrier on or near that frequency. Turn the receiver's BFO on (or switch the receiver to the USB position), and tune in the carrier until you get a zero beat (the tone pitch gets too low to hear). Set the receiver's dial to read exactly 3.500 MHz. If your receiver has both "main-tuning" and "bandsread-tuning" controls, set the bandsread control to 3.500 MHz and tune the main-tuning knob for zero beat. Leave the BFO on, or leave the MODE switch in the USB position.

What if you don't hear anything around 3.500 MHz? That means that the oscillator is not working. The problem may be improper wiring, including solder bridges or cold-solder joints; an incorrect component-value; a bad crystal; a faulty component; or a dead battery. (I had a lot of trouble getting my oscillator to start up until I replaced the crystal.)

To check the amplifiers and modulator, plug a piece of wire about 20 feet long into the input of the converter. You should hear a loud buzz at VLF frequencies, and possibly well up into the LF band. You may also hear a lot of carriers. If you don't, there is something wrong with one of the amplifier stages. Again, check for improper wiring, an incorrect component value, a faulty component, or a weak or dead battery. Once you are sure that the circuit is working, you're ready to put up an antenna.

Not just any old wire lying on the ground, or thrown up into a tree, will work well at VLF and LF. Unless the right kind of antenna is used, all you'll hear is an overwhelming conglomeration of interference including AC-line buzz and cross-modulation products from the AM broadcast band.

Antenna systems

The importance of a good ground system cannot be overemphasized for low-band reception. Preferably, the house utility-ground should be used; look for a pipe running down the electric meter into the ground. Cold-water pipes will also work fairly well, but don't try hot-water or heating-system pipes. The ground connection should be made to the shield at the antenna-end of the coaxial cable used to connect the antenna to the converter.

For low-band reception, a loop antenna generally works best. It should be in a vertical plane, since VLF and LF signals tend to be vertically polarized. Such an antenna can easily be tuned to the desired frequency, and has a narrow bandwidth, which is a necessity for rejection of noise and cross-modulation distortion products. There are two possible configurations for a loop-type receiving antenna: the open loop and the ferrite loopstick. Both are shown schematically in Fig. 4.

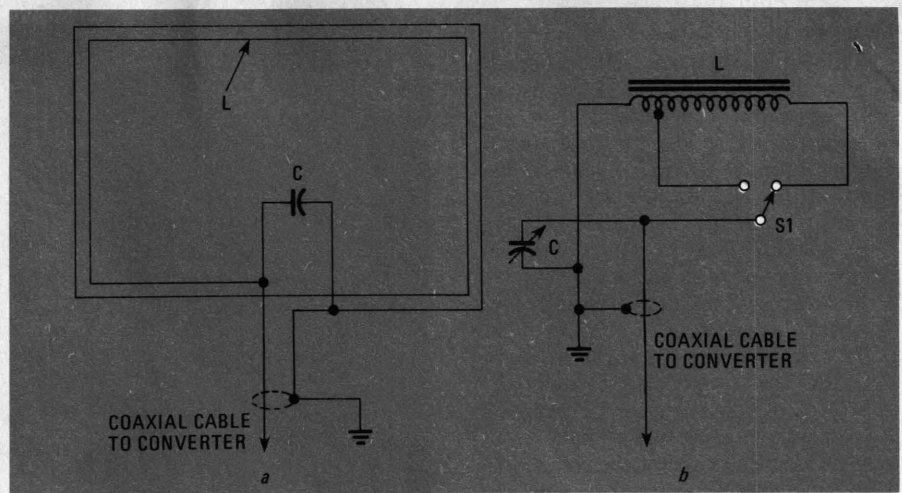


FIG. 4—OPEN-LOOP ANTENNA is shown in a; ferrite loop at b.

An open loop

An open loop consists of eight to twelve turns of insulated wire, mounted in the shape of a square, rectangle, or circle against a non-metallic wall, or between two non-metallic supports. The loop should have a radius of at least two feet. The larger the enclosed area of the loop, the better the signal pickup will be; reception also improves with the number of turns used. Excessively large loops, however, will pick up a great deal of AC buzz and appliance noise, so there is a practical maximum size-limit. That limit will depend on the amount of noise in your area, and will have to be determined by experimentation. The loop should, of course, be placed as far away from AC wiring as possible.

The loop should be tuned to, or close to, resonance at the desired frequency by means of a capacitor connected in parallel with it. The resonant frequency for a given capacitance C depends on the loop inductance L. Assuming that the loop is at least several feet away from large metallic objects, its inductance can be found from the formula:

$$L = \frac{N^2 \sqrt{A/\pi}}{9000}$$

where L is in millihenries (mH), A is the enclosed area of the coil in square inches, and N is the number of turns. If the coil is a perfect circle of radius r inches, then:

$$L = \frac{N^2 r}{9000}$$

Once you know the inductance of the loop, you can determine the amount of capacitance needed for a given frequency f, according to the formula:

$$C = \frac{1000}{4\pi^2 f^2 L}$$

where L is again in mH, f is in kHz, and C is in microfarads (μF).

The capacitance values generally required to tune an open loop to resonance range from about 100 pF at 535 kHz to as much as 0.5 μF for small loops at VLF. It is impractical to use variable capacitors

PARTS LIST

All resistors 1/4-watt, 5%, unless otherwise specified

R1, R3, R4, R6, R7—10,000 ohms

R2, R5, R8, R12—120,000 ohms

R9—R11—100 ohms

Capacitors

C1—C4—0.1 μF , ceramic disc

C5—0.001 μF , ceramic disc

C6—0.01 μF , ceramic disc

C7—2 pF, ceramic disc

Semiconductors

Q1—Q3—2N2484 or equivalent

Q4—2N5951 or equivalent N-channel

JFET

XTAL1—3.500 MHz, parallel-resonant, 20 pF (if necessary, a crystal of another frequency may be used)

L1—1 mH

J1, J2—RCA phono jack, chassis mount

S1—SPST toggle switch

B1—9-volt transistor battery

Miscellaneous: "experimenter board" (see text), wire, battery clip, metal enclosure, antenna materials (see text)

for tuning an open loop over a wide range of frequencies, simply because they don't make them big enough! That means you will have to switch among several fixed capacitors to obtain wide-band coverage. The easiest way to do that is to use a decade capacitance-box. Ceramic, Mylar or mica capacitors are best; don't use electrolytics.

It is important that the tuning capacitors be placed at the antenna, and not at the converter end of the feed line because the line capacitance will have a detrimental effect on selectivity toward the upper end of the LF band. That will allow more cross-modulation distortion from the AM broadcast-band and give the illusion of LF signals where there are none.

A ferrite loopstick antenna

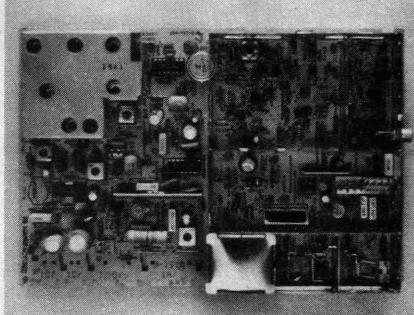
The inconvenience of having to switch among fixed capacitors for frequency adjustment can be overcome by the use of a ferrite loopstick antenna. (That's the kind of antenna that you find in most

continued on page 102

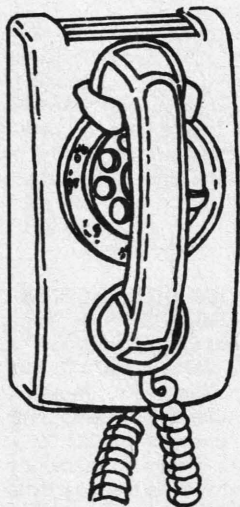


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small, transistorized AM radios.) The arrangement shown in Fig. 4-b can be tuned from 10 to 60 kHz or from 60 to 360 kHz, depending on the amount of inductance that is switched in by S1. The ferrite rod should be as big as one as you can get—5 × 3/8-inches is a good size to work with. The capacitor can be any common 365-pF or 400-pF receiving-type air variable. The exact number of turns needed on the ferrite rod, and the position of the tap, will depend on the size and permeability of the ferrite rod used. The inductance required to tune 10 to 60 kHz is 685 mH, and the inductance required to tune 60 to 360 kHz is 19 mH. Winding data is generally given with ferrite rods; the values can also be found by trial and error. Use fine enameled wire, such as No. 30 or 32, for winding.

One advantage of the ferrite loopstick is that it can be easily positioned to null out man-made noise. Heating pads, electric blankets, vacuum cleaners, and hair dryers are notorious for their ability to saturate the electromagnetic spectrum with noise well up into the VHF range; at VLF and LF, they can be devastating! By turning the loopstick in both the vertical and horizontal planes, that kind of noise, as well as the AC buzz that always seems to hinder VLF reception, can be nearly eliminated. A ferrite loop will also provide you with a sharper degree of selectivity than an open loop. That insures still more noise immunity.

The principal disadvantages of the ferrite loopstick are lower signal-capturing ability because of its smaller physical size, and the difficulty of locating a rod big enough to wind an inductor of 685 mH. A military surplus shop is a good place to look for large ferrite rods. Electronics outlets also may carry them.

My preference is the open loop, primarily because of its superior signal-pickup and ease of construction. At times, someone runs a hair dryer or vacuum cleaner, but my ham-band receiver has an excellent noise blander, and that is satisfactory under all but the worst conditions.

Dealing with noise

The tuned-circuit resonance provided by the parallel capacitors, both with the open loop and the ferrite loopstick, give some noise reduction because they cause the antenna to have a narrow-band response. Both antennas can be mounted on azimuth/elevation bearings (although that's very difficult with a large open loop), and positioned for minimum noise. If it cannot be rotated for a null in the noise, the open loop can be used with a noise-cancellation circuit that often works remarkably well. Such a circuit is

shown schematically in Fig. 5.

Both the loop itself, and the noise-pickup wire P, receive impulse noise generated by the surrounding appliances and utility wires. But the loop receives far

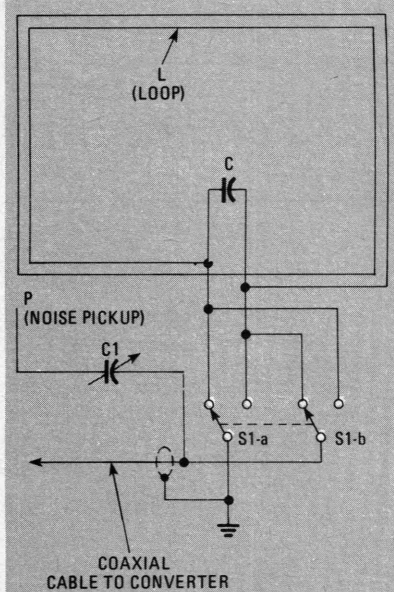


FIG. 5—OUT-OF-PHASE signals from wire "P" and loop "L" provide noise-cancellation.

more of the desired signal-energy than wire P. Noise cancellation is obtained by first finding the loop connection (via S1) that causes the noise picked up by the loop to be out of phase with the noise from P. Capacitor C1 is then adjusted until the noise inputs from P and from the loop are equal in amplitude. At that point, there will be a sharp drop in the receiver noise, but the signal level will not be affected.

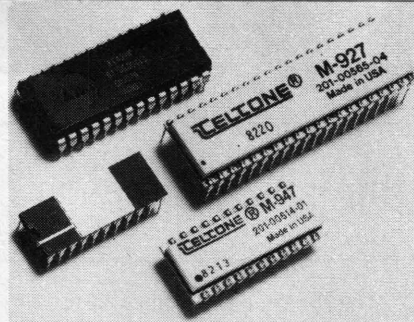
It may be necessary to experiment with various lengths of wire for P. If your receiver has a signal-strength indicator (S-meter), the task can be simplified by noting the noise level with the loop only, and then adjusting C1 for an equal noise level using P only. If the noise level is always lower when using P alone than when using the loop alone, you will have to lengthen P. If the noise level is higher with P alone, you will have to shorten it. Ideally, C1 should be at the middle of its tuning range when the noise pickup is equal from both antennas.

Once the two antennas are tied together, there will be some mutual-coupling effects that will require you to change the setting of C1 somewhat. However, the above method should bring you close to the required length for P and the required setting for C1.

The noise-cancelling circuit just described usually works better at VLF than at LF-or-higher frequencies, and is more effective against some kinds of noise than others. It will be of little value in reducing AM-broadcast cross modulation or atmospheric static. Still, I've found that it allows much better reception in the VLF range, as compared with an open loop by itself.

R-E

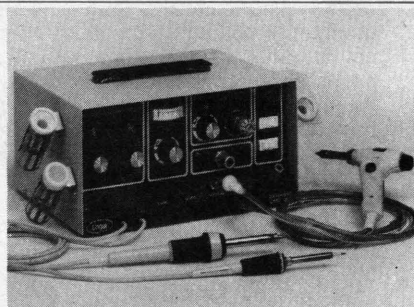
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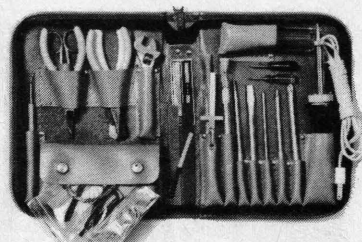
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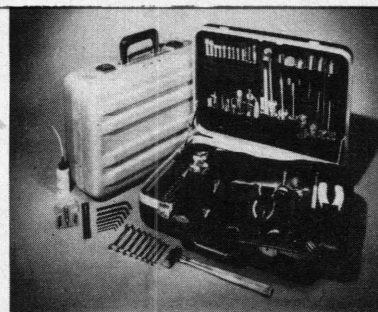


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